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Engineering Aspects of the TH Microwave Radio Relay System

By J. P. KINZER and J. F. LAIDIG

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This paper gives a general description of the TH system, which is a long-haul system operating in the 5925–6425 mc band. It discusses the considerations governing the major design decisions, which lead to an FM system of eight channels in each direction, six working and two protection. Each channel has a 10-mc potential baseband width. The paper continues with a discussion of noise considerations, primarily from the viewpoint of telephone transmission. Fluctuation noise and modulation noise are the principal sources, with interference between radio channels a minor factor.

I. INTRODUCTION

The TH radio relay system is the most recent in the succession of radio facilities developed for use in the Bell System communication network. A point-to-point radio system was used as part of the commercial telephone network as early as 1920.[1] This and subsequent point-to-point radio installations were special cases using a single-hop, narrow-band system to carry one or a few telephone conversations across terrain where installation of wires or cables was impracticable. The use of radio relay as a general purpose facility in the Bell System network was inaugurated with the installation of the experimental TD-X system between New York and Boston in 1947. From this initial 220-mile route, the radio network had expanded by the end of 1960 to over 44,000 route miles carrying over twenty-nine million telephone circuit miles and 79,000 television channel miles.

The major portion of this network is provided by the long-haul broadband TD-2 system operating in the 4000-mc common carrier

band. A relatively small amount is provided by short-haul feeder systems. Many of the TD-2 routes are now fully equipped and no more radio channels can be added within the assigned frequency band. The congestion is particularly acute where two or more routes cross or converge on cities. Additional circuit needs in these areas can be met only by the use of a different frequency band.

The TH radio relay system has been developed to provide the additional circuits. It is a broadband long-haul radio system operating in the common carrier band between 5925 mc and 6425 mc.[2] The basic design objective is to provide 4000-mile telephone and television circuits with transmission quality and reliability commensurate with the present system objectives. At the same time, maximum utilization of the available frequency spectrum is imperative, and the use of as much as possible of the existing TD-2 plant and engineering effort is economically desirable. Along with these requirements is the need to provide flexibility to meet changing conditions of usage, both in layout of routes and in types of signals to be carried. This paper gives a brief description of the system and discusses the over-all engineering aspects which led to the final system. Subsequent papers discuss in greater detail the design of the various component parts of the TH system.

II. THE TH SYSTEM

The TH system is shown in block diagram in Fig. 1. Each broadband circuit consists of an FM transmitter, a transmitting entrance link,

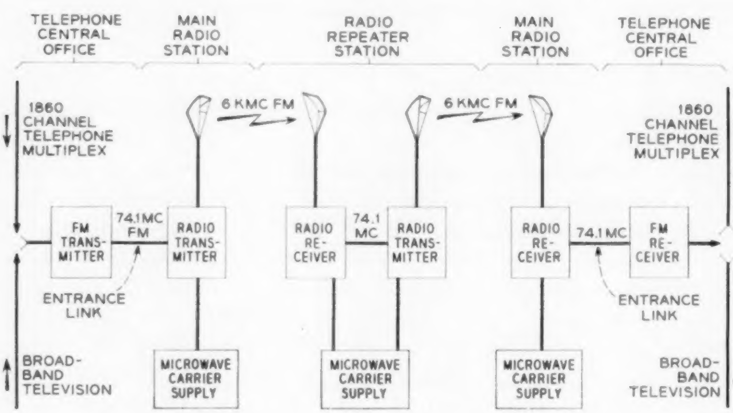

Fig. 1 — Block diagram of the TH system.

one or more radio repeater sections, a receiving entrance link and an FM receiver. The FM transmitter produces a 74.1-mc signal which is frequency modulated by the incoming baseband signal. This frequency-modulated signal is fed to a radio transmitter over a transmitting entrance link. The radio transmitter heterodynes the IF signal to the desired microwave frequency and amplifies it for radiation to the next station. The radio receiver at the next station heterodynes the microwave frequency back to the 74.1-mc intermediate frequency for amplification and equalization. The output of the radio receiver may then be connected to another radio transmitter for transmission farther down the route or may be connected through a receiving entrance link to an FM receiver. The FM receiver demodulates the signal and delivers the baseband signal to the terminating equipment.

Eight two-way, broadband radio channels and two two-way, narrowband radio channels (called the auxiliary channels) are provided in the assigned band as shown in Fig. 2. In each station, all the radio transmitters are grouped in one half of the assigned band and all the radio receivers in the other half of the band. Six of the broadband channels are used as working channels with the other two used as protection channels against equipment failures or fading. The auxiliary channels are located in the guard spaces between the broadband transmitters and receivers and at the edges of the band. These are used to provide circuits for voice order wires, for the automatic protection switching system, and for transmission of alarms from unattended stations. The two auxiliary channels in each direction are connected to provide a single transmission circuit with frequency diversity.

A fully equipped repeater station contains twenty radio receivers and twenty radio transmitters, each requiring a microwave beat oscillator signal. Rather than providing forty individual oscillators, a common carrier supply is used for all the receivers and transmitters in a station. Starting from a highly frequency-stabilized crystal oscillator operating at 14.8259 mc, the common carrier supply generates the 2nd, 4th, 408th and 425th harmonics to provide two VHF signals at 29.65 mc and 59.30 mc, and two microwave signals at 6049 mc and 6301 mc. The appropriate one of the ten beat oscillator frequencies shown on Fig. 2 is provided by either the 6049-mc or 6301-mc signals or is generated in the receiver or transmitter by combining one of the microwave signals and one of the VHF signals. Since the failure of the common carrier supply in a station would be catastrophic, it is supplied in duplicate with automatic switching to the standby unit when necessary.

The baseband signals to be carried by the TH system are shown in

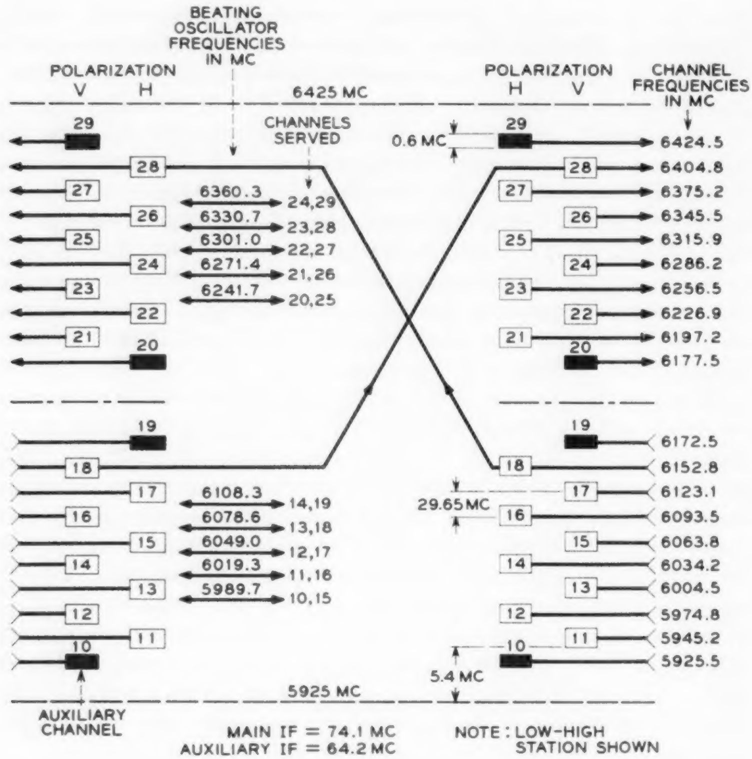

Fig. 2 — Frequency allocations and polarizations for the TH system channels.

Fig. 3. The telephone signals are derived by single-sideband multiplex equipment of the type used in the L-3 coaxial cable system.[3] The telephone signal for TH differs from the standard L-3 telephone signal only in the amount of pre-emphasis. High-definition theater television occupying the entire 10-mc baseband may also be transmitted.

III. SYSTEM DESIGN CONSIDERATIONS

The TH system as described above is basically similar to the existing TD-2 system but contains many refinements. This is accomplished through the use of newly developed components and through applying the experience gained from the TD-2 system. The design departs from that of the older system wherever an improvement is indicated and

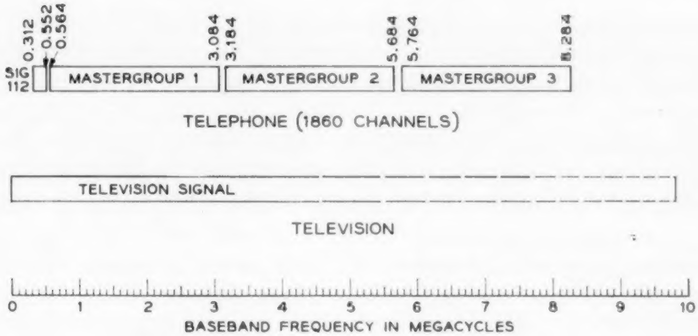

Fig. 3 — The baseband signals of the TH system.

follows that of the older system only if a review of the various possibilities showed that approach to be best.

3.1 *Antenna System*

A large portion of the expense of any radio relay route is the cost of repeater sites, access roads and towers to support the antennas. Any new radio system should be designed to utilize the repeater sites and towers of older systems if at all possible. Furthermore, the same antennas should be used for as many systems on any route as possible. The cost penalty for using separate antennas for each system involves provision of heavier, more expensive towers and additional waveguide runs as well as the additional antennas themselves. In anticipation of the use of TD-2 (4 kmc), TH (6 kmc) and possibly TJ (11 kmc) systems on the same route, the horn-reflector antenna was developed.[4,5] This antenna, with its circular waveguide feed, can transmit cross polarized signals in all of these bands. Polarization is used to help separate two signals which are too close together in frequency to be separated completely by practicable filters. The cross polarization discrimination of a repeater section, including the antennas, round waveguides and system combining networks, is expected to be at least 25 db for the TH system. The performance of the antennas is one of the factors permitting the close spacing of adjacent channels as shown in Fig. 2.

The horn-reflector antenna is designed to have approximately the same gain in the 4-kmc band as the delay lens antenna originally used with TD-2. This in turn sets the antenna gain at 6 kmc as 43 db. The average spacing of TD-2 repeaters is about 30 miles. The 6-kmc path loss between isotropic radiators spaced 30 miles apart is 142 db. Thus

the requirement for maximum utilization of existing TD-2 plant determines the loss between the transmitter output and the receiver input of the TH system. With an allowance of 8 db for losses in the waveguide components, this is 64 db.

The connections to the antenna are shown diagrammatically in Fig. 4, which shows only one side of the TH repeater station. The system combining networks[6] at the bottom of the round waveguide accept signals

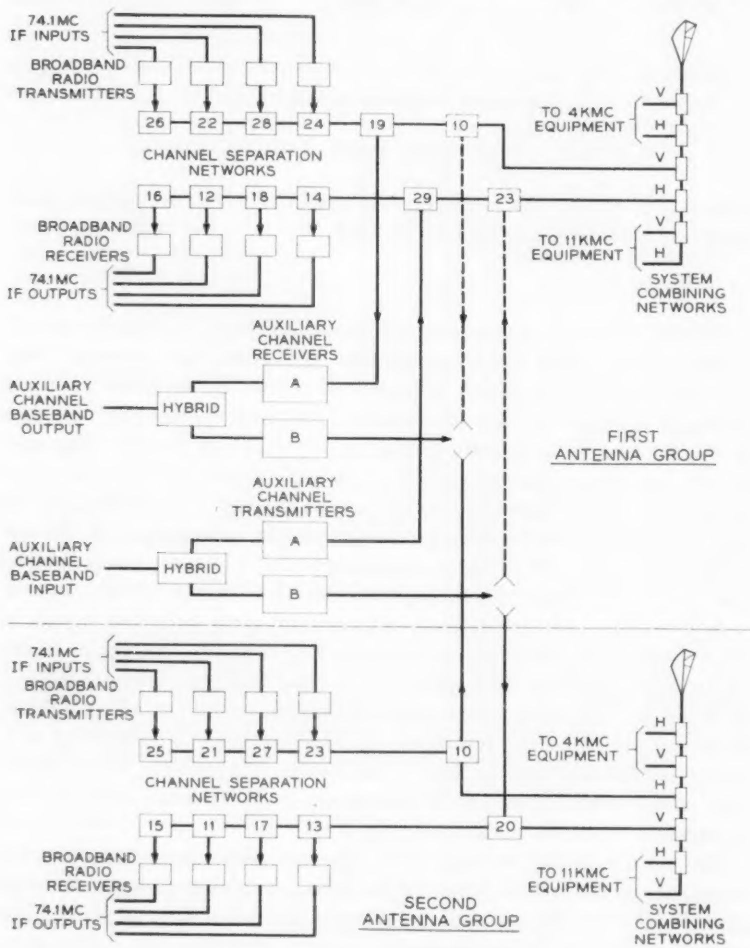

Fig. 4 — Diagram of antenna connections.

in two rectangular waveguides for each frequency band and combine them into cross polarized signals in the round waveguide. The individual channels of the TH system are combined in the tandem connected channel separation networks.

The use of the same antenna for transmitters and receivers permits improved reliability against failures in an antenna or its round waveguide on a heavily developed route, since failure of any one antenna does not shut down all radio channels. The initial installation of TH normally consists of one working and one protection broadband channel. By the standard growth plan, these are channels 4 and 8 (channel 4 designates the alternate sections of channels 14 and 24 required to make up a continuous transmission path, etc.). As additional broadband channels are required, a second pair is added to provide a second working and a second protection channel. These may be either the other two even-numbered channels or a pair of odd-numbered channels (normally 3 and 7). If the first four are all even-numbered channels, they operate on a single pair of antennas as shown in the upper part of Fig. 4, and the installation of the second pair of antennas may be deferred until an odd-numbered channel is required. Where the second pair of antennas is available initially, as when TH is added to a TD-2 route already equipped with horn reflector antennas, odd-numbered channels are used for the second pair to utilize the added reliability of the extra antennas. Additional channels are then added as needed until all channels are equipped as shown in Fig. 4.

Normally, channels 1 and 5 are equipped last to leave two slots in the band which may be used by nearby light-route equipments as long as possible.

3.2 *Frequency Allocation*

The frequency allocation plan of TH provides maximum utilization of the assigned band consistent with the transmission objectives. Interference between transmitters and receivers in the same station is minimized by grouping all the transmitters together in one half of the band and all the receivers together in the other half, as shown in Fig. 2. This, together with the use of cross polarization, permits the guard bands to be reduced to one at the center of the band between the transmitters and the receivers, and one at each edge of the band between the TH signals and any services in adjacent bands. Because of the excellent back-to-back discrimination of the horn-reflector antenna, the same frequency may be used to transmit in both directions at each station, and the full band may be utilized in both directions.

As shown in Fig. 2, each channel is shifted in frequency (or frogged) by 252 mc as it passes through a repeater station. Frequency frogging is necessary to prevent feedback from the repeater output into the input. In TH, if a transmitter and a receiver are operated at the same frequency in the same station, the loss between the transmitter and the receiver has to be greater than 118 db to keep interferences at a tolerable level. Only ideal locations with tall towers and practically no nearby reflective surfaces can be made to meet such a requirement.

Minimum channel spacings are determined by adjacent channel interference considerations. With low index frequency modulation, the interference increases sharply when the channel spacing is reduced to the point that the second-order sidebands of one channel overlap the first-order sidebands of the other. Thus the minimum channel spacing is set at three times the top baseband frequency. For the 10-mc nominal top baseband frequency of TH, the channel spacings must be about 30 mc. These considerations determine the general frequency allocation plan of TH. The exact choice of frequencies is discussed in connection with the common carrier supply, later in this paper.

3.3 *Type of Modulation*

The type of modulation to be used in a system is chosen to give the best compromise between noise and distortion performance on the one hand and practicable design limitations on the other. For example, pulse modulation systems can be made to give excellent noise and distortion performance through many repeaters in tandem, but they are not suitable for use in the TH system because of the large bandwidths they require. High index frequency modulation is unsuitable for the same reason since it too trades bandwidth for noise performance. At the other extreme, single sideband suppressed carrier transmission makes the most efficient use of radio spectrum and of transmitter power, but it requires a high degree of amplitude linearity in the amplifiers to reduce distortion. At the present state of the art, 6-kmc power amplifiers of the required degree of linearity would be prohibitively large and expensive, if they could be built at all. The compromise chosen for TH is low deviation frequency modulation. This requires a relatively narrow band (approximately three times the top baseband frequency) and will tolerate compression in the amplifiers.

The optimum frequency deviation to be used is related to many factors, among which the more important are the transmitter power, the noise figure of the radio receiver, the system delay distortions, and the baseband signal to be transmitted. At the time TH was being

planned, a traveling-wave tube in a reasonable size with 5 watts output became a practical possibility. This was chosen as the output stage for TH, and the tube was developed as the Western Electric Type 444A.[7,8] New modulator crystals and the use of an isolator in the modulator input combine to give a receiver noise figure of 10 db. All components of TH are designed for low delay distortion by using the most stable elements available in circuit configurations having minimum delay distortion. The traveling-wave tube amplifier is designed to be essentially flat over the entire 500-mc TH band with only minor tuning adjustments required for each channel. The IF amplifier uses factory adjusted, fixed-tuned interstage circuits designed to minimize effects of changes in tube parameters. The radio repeater room is temperature controlled to minimize temperature effects on waveguide filters. When these parameters are considered in terms of the 1860 telephone channel load, the optimum frequency deviation is found to be ± 4 mc peak, which is about the same as that used in TD-2.

3.4 *Type of Repeater*

Three general types of radio repeaters might be used. These are the baseband repeater, in which the signal is demodulated to baseband and remodulated on the radio frequency at each repeater; the IF repeater, in which the modulated radio frequency signal is heterodyned to an intermediate frequency for amplification and equalization and then heterodyned back to radio frequency without demodulation; and the RF repeater, in which the signal is amplified and reradiated with only the frequency shift required for frogging. An RF repeater may be either a single radio channel repeater in which the radio channels are separated and amplified in individual amplifiers, or a multichannel repeater in which all radio channels in one direction are amplified in a single amplifier.

The baseband repeater provides the highest degree of route flexibility by making the baseband signal available at all repeater points along the route. However, in a long-haul system such as TH, the large number of modulators and demodulators connected in tandem would require each modulator to meet extremely stringent distortion requirements. The multichannel RF repeater is not suitable for applications where polarization is used to separate the signals of adjacent radio channels. This leaves only the single channel RF repeater and the IF repeater as possible choices for TH. The latter was chosen because of the greater ease of separating and equalizing channels and of switching between working and protection channels at IF.

The use of IF-type repeaters leads to the use of the intermediate frequency as a common interconnection between the various portions of the system. The output of the FM transmitter and the input of the FM receiver are made to operate with the same IF signal as is used in the radio repeaters. Entrance links, when required to connect widely separated radio equipment and multiplex terminals, are also operated at IF. Good route flexibility results, since all route reassignments or additions are made by IF interconnections.

Tone interference due to leakage of the beat frequency oscillators into the radio channels can be minimized by the proper choice of the intermediate frequency. If the IF is made equal to an odd multiple of half the channel spacing, the leakage tones fall midway between two channels and are least disturbing to both. Past experience with other microwave systems has indicated that an IF of about 70 mc is a good compromise in the IF amplifier design between electron tube input conductance and broadband coupling network design. This leads to the use of $2\frac{1}{2}$ times the channel spacing (74.1 mc) for the TH intermediate frequency. This IF is not compatible with the 70-mc IF of TD-2. However, regular use of direct IF interconnections of the two facilities would be extremely unlikely, since the use of the 10-mc bandwidth of TH to carry the 2.25-mc to 4-mc signal of TD-2 would be economically wasteful and since the use of the TD-2 system to carry the wider signal of TH is impossible. For emergency restoral of service, the wider IF bandwidth of TH will permit transmission of the offset TD-2 signal although at the expense of possible interchannel interferences of the type discussed later as "tertiary interference".

3.5 *Common Microwave Carrier Supply*

Forty microwave beat oscillator signals are required in a fully equipped radio repeater station. The use of a common carrier supply rather than individual supplies reduces the active equipment required, reduces maintenance efforts, eases stability requirements, and permits better control of tone interferences.

The frequency stability objective of the TH system (derived from interference and distortion considerations) is that at no point in a 4000-mile circuit shall the transmitted frequency be in error by more than 280 kc (one per cent of the transmitted bandwidth). Of this tolerance, 100 kc is allocated to the FM transmitter with the remaining 180 kc available for errors in the carrier supplies of 133 tandem radio repeaters. If individual supplies were used, the 180 kc would have to be divided among 266 supplies. Assuming random addition of the errors,

this permits about 11 kc error per supply or less than 2 parts per million. However, if a common carrier supply is used for both the receiver and the transmitter of a channel, only the error in the shift frequency need be considered in that station. This is true whether the transmitter and receiver beat oscillators are derived from different harmonics of the same oscillator or from the combination of a microwave oscillator and a shift oscillator. This reduces the error sources in a 4000-mile route to two 6-kmc carriers in the end stations in tandem with 132 shift oscillators operating at 252 mc. Again assuming random addition and that the stability of the shift oscillators is the same as that of the carrier supplies at the ends of the route, the permissible error becomes 20 parts per million. The actual oscillators used in TH have approximately an order of magnitude better frequency stability than required above.

The frequency stability discussed above can be attained either with a common carrier supply of the type used in TH or with carrier supplies individual to the channels as is done in TD-2. In either case, the stability requirement calls for use of a crystal oscillator as the basic frequency-determining circuit, with frequency multipliers to generate the microwave frequency. In a fully equipped station, individual channel supplies of the TD-2 type would require about three times as many electron tubes as does the common carrier supply. Since the number of electron tubes can be taken as a rough index of the total number of components required and of the failure rate, the common carrier supply should be considerably less expensive and require considerably less maintenance effort in a fully equipped station. In the minimum size station (two two-way channels) the individual carrier supplies would have fewer components, but since the cross-over point occurs at only three two-way channels, the advantages of the common carrier supply outweigh any initial cost advantage for such minimum routes.

The beat oscillator (BO) frequencies are shown in the center of Fig. 2. The upper five radio channels in each half of the band (including the upper auxiliary channel) use BO frequencies lower than the signal frequencies, while the lower five channels use BO frequencies higher than the channel frequencies. Since the IF has been chosen to make the BO frequencies fall midway between channels, each BO frequency serves two channel assignments, and only five BO frequencies are required in each half of the band, or a total of ten in each station. These ten frequencies are derived from four frequencies generated in the common carrier supply. The center BO frequency in each half of the band (6049 and 6301 mc) may be used directly or may be shifted up or down by one or two channel spacings (29.65 or 59.30 mc). These four

frequencies are distributed from the common carrier supply as needed to the transmitters and receivers, where individual carrier shift modulators and microwave filters generate and select the desired frequency.

Tone interference requirements, which are very stringent in the TH system, are discussed later in this paper. They lead to severe requirements on the microwave filters following the shift modulators. However, if the basic frequencies fed to the shift modulators are all harmonics of the same oscillator, the leakage tones will all fall exactly on the BO frequencies or between channels where the requirements are most lenient. Another argument for a coherent common carrier supply (one in which all frequencies are harmonics of a single oscillator) is that of maintenance. If independent oscillators are used, maintenance facilities must be provided for measuring and adjusting all the various oscillator frequencies to close tolerances; whereas in the coherent supply, only one oscillator frequency need be measured and controlled.

A coherent common carrier supply, although desirable, is not necessarily feasible. A number of independent requirements must be met, and the oscillator frequency is the only variable which may be adjusted to satisfy them all. The basic oscillator frequency must be an integral submultiple of the channel spacing. The two microwave frequencies generated in the carrier supply must be obtainable from the basic oscillator by simple multiplication chains and a minimum of shift modulators. Multiplication factors must be no more than two or three to aid in rejection of spurious harmonics. A center guard band must be provided. The BO frequencies must be symmetrical about the center of the assigned band (6175 mc). And finally, the channel spacing must be approximately 30 mc. Fortunately, a reasonable solution was found for TH as shown in Fig. 5. The oscillator frequency is 14.82593 mc, which sets the channel spacing at 29.65 mc and the intermediate frequency at 74.13 mc. This also determines the exact broadband channel frequencies as shown in Figs. 2 and 5.

In comparing the reliability of the common carrier supply with that of individual channel carrier supplies, both single channel failures and complete route failures must be considered. The failure of any one tube in an individual channel carrier supply will cause the failure of one radio channel. As a result, failure of one or two radio channels due to carrier supply failures is not particularly rare. However, failure of an entire route due to carrier supply failure is almost unthinkably improbable. A common carrier supply provided with one complete standby generator with automatic switching reduces the single channel carrier supply failures to negligibility but does introduce the possibility of

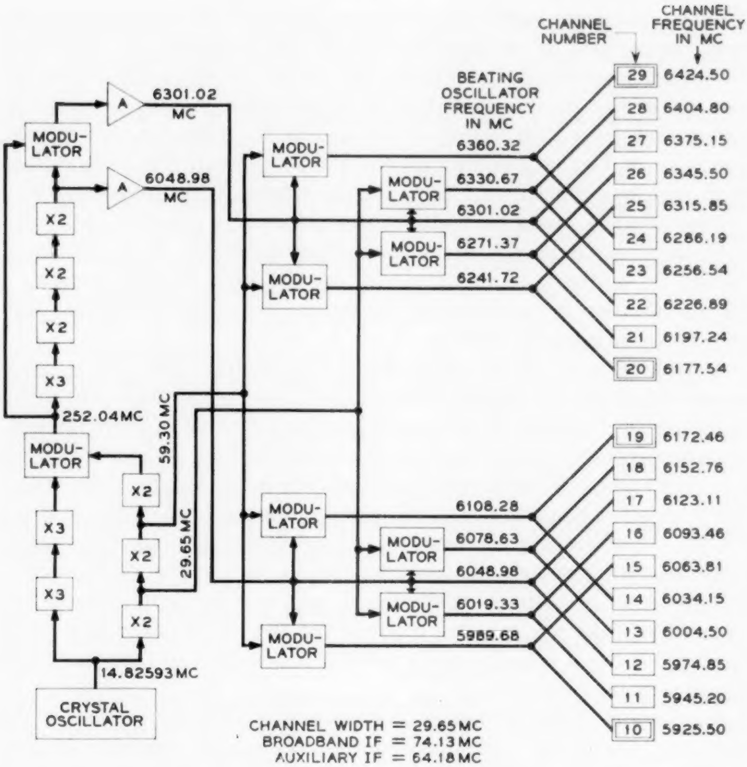

Fig. 5 — Generation of beat-oscillator frequencies from single crystal oscillator.

complete route failure. A study based on anticipated tube and component failure rates gave an expected route failure rate of from 0.007 to 0.015 failure per year per 1000 miles. The addition of a second standby generator would reduce this rate by several orders of magnitude. However, this predicted failure rate with one standby is comparable to failure rates in other currently used broadband systems, and the second standby generator does not appear to be warranted.

3.6 *Power System*

A high reliability radio system, must, of course, have a highly reliable power source. In other radio systems, reliability is attained by operating

the electronic equipment from batteries which in turn are kept charged from commercial ac or from emergency engine-alternators. However, the traveling-wave tubes of TH require voltages of up to 3000. Both the size of battery plant required and the personnel hazard involved in distributing such high voltages rule against batteries for this service. The safer and more practical approach is to generate the high voltage dc in each bay containing a traveling-wave tube. While individual high voltage supplies could be built for operation from either ac or dc inputs, an ac input supply is simpler and more reliable. If a reliable, or "firm," ac source is to be developed for the high voltage supplies, a logical extension is to operate all electronic components from local supplies powered from the firm ac.

A bank of from four to seven generators provides 230-volt, 60-cycle ac power for the TH system. These generators are normally driven by ac motors operating from commercial ac. In the event of power failure, dc motors mounted on the same shaft are switched on, and the ac output is uninterrupted. If the power failure persists long enough for the emergency diesel engine-alternator to pick up the ac load of the station, the ac motor is again activated, this time operating from the engine-alternator. After restoral of commercial ac, the ac motor is automatically switched back to normal. While the dc motor is normally intended to operate only the few minutes required for the engine-alternator to be activated, several hours' battery reserve is provided in case the diesel engine fails to start.

A minimum of two generators is operating at all times to provide the firm ac. The regular and standby common carrier supplies, the two auxiliary channels, and other critical loads are divided between these two machines so that failure of either can not cause total failure of the station. As more channels are installed, generators are added to a maximum of five required to carry a fully equipped TH station. Loads are distributed among these generators for maximum reliability. In addition to the normally working generators, a running spare and a nonrunning spare are provided in each station. The running spare is automatically switched in place of any working machine that fails. The nonrunning spare may be manually patched to replace any of the running machines to permit maintenance or repair without loss of reliability.

3.7 *Auxiliary Channels**

Communication facilities between the various stations of the system are necessary for voice communications between maintenance personnel,

* Material contributed by A. V. Wurmser.

transmission of alarm indications from unattended stations, and operation of automatic protection switching. These services were originally provided for TD-2 routes by wire line networks. However, radio repeater stations are normally located on the tops of hills while cable or wire line routes tend to follow valleys, so that expensive and difficult-to-maintain spur lines are usually required to connect the individual radio stations to the wire networks. Many of the more recently installed TD-2 routes are equipped with light-route radio systems to provide the maintenance circuits. The TH auxiliary channels are provided as an integral part of the system, to carry maintenance circuits.

The auxiliary channels must be inexpensive but reliable: inexpensive so they can compete economically with wire lines or separate light-route radio systems, and reliable so they will be operable when needed to transmit vital control information for the broadband circuits. These needs, together with the requirement that the auxiliary channels operate in a frequency band assigned to common carrier use, resulted in the decision to integrate the auxiliary channel into the TH system and make maximum use of the performance and reliability of the high gain antenna system and the common carrier supply.

The auxiliary channels must be protected against transmission failures due to either equipment failures or fading. To reduce interruptions of this nature to a minimum, twin channel transmission on a frequency diversity basis is used. Statistical data indicate that, for frequency diversity to be as effective as possible, the two channels carrying duplicate signals should be at least 160 mc apart. The location of the auxiliary channels at band edges and band center, as shown in Fig. 2, gives nearly 250 mc frequency diversity.

The type of modulation used on the auxiliary channels is determined primarily by the need to demodulate to baseband at each station. A study indicated that either AM or FM is practical but that the AM system would be considerably more economical because of the complexity of FM terminals, particularly in comparison to double sideband AM terminals.

In addition to the maintenance circuits of TH, the auxiliary channel accommodates the TD-2 maintenance circuits on routes common to both systems. Voice communication and alarm and control channels to intermediate stations can be common to TD-2 and TH. These functions require two two-way telephone channels. Automatic protection switching control signals, however, require different channels for TD-2 and TH because of the frequent differences in the locations of the switching terminals as well as the dissimilarity between the switching signals.

The TD-2 switching signals are modulated tones and require one two-way telephone channel. The TH switching signals consist of 16 single-frequency tones in each direction and occupy a bandwidth of 16 kc. One additional voice channel is provided for use where needed for direct communication between terminals or other purposes. Thus the minimum capacity provided by the auxiliary channel is equivalent to four telephone channels plus a 16-kc band for TH automatic switching tones.

The design also allows expanding the auxiliary channel to provide a small number of short-haul telephone channels in conjunction with the basic four channels. The ON carrier system, in which the basic group consists of four telephone channels, lends itself to this use. The 16 TH switching tones are spaced 1 kc apart between 20.5 kc and 35.5 kc. The ON operating frequency range begins at 40 kc, just above the tones, and terminates at 264 kc. With this future possibility in mind, the top baseband frequency was set at 300 kc, resulting in a required radio frequency bandwidth of 0.6 mc, for a double sideband AM system.

The transmitting power for the basic auxiliary channels can be obtained directly from the common carrier supply, but if any short-haul telephone channels are added, a traveling-wave tube (TWT) amplifier is required in the transmitter output. This conclusion is based on noise and crosstalk requirements for a 10-repeater system, this length being considered typical of short-haul telephone service. Since the TWT amplifier would add considerable cost to systems that do not need this service, the development of that portion of the circuit has been deferred. The basic radio channel circuits are designed to handle a 300-kc baseband signal, and space is provided in the radio terminal bays for future installation of the TWT equipment if the need develops.

Studies of performance of the basic system showed that objectionable interchannel cross-modulation can occur between the TH switching tones and the four voice channels. To avoid this problem, the basic four voice channels are assigned to the ON group operating in the 80- to 96-kc baseband range.

The auxiliary channels are operated on the same antenna system with the broadband channels. In the fully equipped system, maximum equipment diversity is obtained by connecting one auxiliary channel receiver and one auxiliary channel transmitter to each antenna as shown by the solid lines of Fig. 4. If, during the early growth stages of a route, only two antennas are provided per station, the auxiliary channel B transmitter and receiver must be connected to the first antenna. This temporary connection is shown dotted in Fig. 4. During this period, the separation between channels 19 and 20 is insufficient, so channel 20 is

temporarily shifted to the channel 23 frequency allocation. When the route grows to the point that the second set of antennas is installed, auxiliary channel B is restored to its normal frequency and antenna connection.

The auxiliary channel separation networks are connected on the antenna side of the broadband separation networks to facilitate the addition of broadband channels after the initial installation without interrupting the working auxiliary channels. The auxiliary channel transmitters are connected with the broadband receivers to relieve some interchannel interferences. These are discussed in Section V of this paper.

IV. NOISE AND EQUALIZATION

In the early stages of system design, the noise objective for the system is customarily divided among various sources as experience and judgment indicate, and efforts are then made to meet the allocations. As the design progresses to final hardware and as natural limitations appear, the original allocations become more or less historical in nature. Accordingly, the noise performance of TH is discussed herein from the viewpoint of expected performance rather than early allocations. The over-all objective for 4000 miles of TH is an rms telephone circuit noise during busy periods from all sources no greater than 39 dba at the 0-db transmission level (TL) in the worst telephone channel.

The sources of noise may be divided loosely into three groups: fluctuation, modulation and interference.

4.1 *Fluctuation Noise*

There are three contributors to fluctuation noise: the amplifiers in the signal path of the radio repeaters, the microwave carrier supplies, and the FM terminals. The radio repeater amplifier noise is generally controlling above 2 mc, the FM terminal noise controls below 0.1 mc, and all three sources contribute in the region between 0.1 and 2 mc. Fig. 6 shows the anticipated values of these sources for 4000 miles of TH radio referred to a 0 dbv point.* The circles are values extrapolated from measurements on the initial installation, without the FM transmitter.

* A 0 dbv point in the TH system is defined as one where a 1 volt peak-to-peak baseband signal corresponds to 8 mc (±4 mc) swing of the FM carrier. The TH baseband circuit impedance is 124 ohms balanced, and a sine wave of 0 dbm is also 0 dbv. The input to the FM transmitter is a 0 dbv point, and the output of the FM receiver is a +8 dbv point.

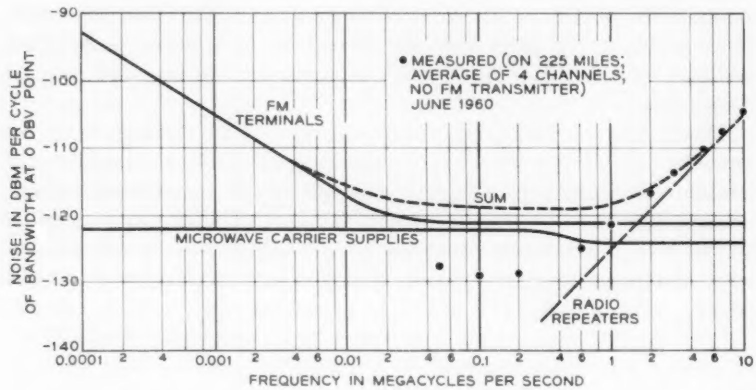

Fig. 6 — Anticipated values of fluctuation noise for 4000 miles of TH radio.

The radio repeater noise in Fig. 6 is based on the normal transmitter power of +37 dbm, received carrier power of −27 dbm (section loss of 64 db) and 10-db repeater noise figure. The 10-db repeater noise figure includes the contribution of the TWT amplifier. The section loss figure is based on study of the 477-mile initial TH installation between Prospect Valley, Colorado, and Salt Lake City Junction, Utah.

The section loss includes channel separation networks, systems combining networks, waveguides and antennas as well as the path loss itself. The section loss for each broadband channel in each of the 15 repeater sections of the initial installation was calculated from the plot plans and building layouts. These losses ranged from 51 db to 66 db, with an rms value of 62.7 db. The channels farthest from the antennas have about 1.3 db more loss than those nearest to it, and will thus have 1.3 db more radio repeater noise. The rms of the actual measured section losses for the channels of the initial installation is 62.4 db.

This particular route was designed with great care to keep waveguide runs as short as possible. Most of the towers are very short and the buildings are very close to the towers. The average future installation will probably include longer waveguide runs. The addition of only 50 feet of rectangular waveguide in each transmitter and receiver connection would increase the section loss by 1.3 db. Thus 64 db is taken as a more representative value for section loss. As an example of how section loss accumulates, Table I gives a detailed breakdown of the extreme channels in a typical repeater section. The importance of short waveguide runs is clearly shown by this tabulation.

TABLE I — SECTION LOSS ANALYSIS, GREEN RIVER TO CHURCH BUTTES

Channel number	26	23
Green River:		
Trans Monitor directional coupler	0.03	0.03
Channel Separation Network	(Ch. 26) 0.40	(Ch. 23) 0.40
Channel Separation Network	(Ch. 22) 0.08	
Channel Separation Network	(Ch. 28) 0.14	
Channel Separation Network	(Ch. 24) 0.14	
Interchannel waveguide	0.55	
Aux Chan Sep Net	0.06	0.06
Indoor Rect waveguide	0.49	0.56
Outdoor Rect waveguide	0.62	0.61
System Combining Network	0.56	0.33
Circular waveguide	0.04	0.04
Sum	3.11	2.03
Path Loss	142.60	142.47
Antenna Gains (2)	−86.16	−86.45
Sum	56.44	56.02
Church Buttes:		
Circular Waveguide	0.24	0.24
System Combining Network	0.56	0.33
Outdoor Rect Waveguide	1.35	1.41
Indoor Rect Waveguide	0.64	0.71
Aux Chan Sep Net	0.06	0.06
Interchannel Waveguide	0.03	
Channel Separation Network	(Ch. 24) 0.14	
Channel Separation Network	(Ch. 28) 0.14	
Channel Separation Network	(Ch. 22) 0.08	
Channel Separation Network	(Ch. 26) 0.40	(Ch. 23) 0.40
Sum	3.64	3.15
Total	63.19	61.20

Green River tower 37.5 feet high, 13 feet from waveguide entrance to building.
Church Buttes Tower 112.5 feet high, 30 feet from waveguide entrance to building.
Path length 31.5 miles.

At the radio receiver input, thermal noise is -174 dbm per cycle of bandwidth (denoted dbm/cbw). With the repeater noise figure of 10 db and the receiver input carrier power of -27 dbm, the repeater noise is $(-174 + 10 + 27)$ or -137 db relative to unmodulated carrier per cbw. The FM deviation constant is such that 0 dbm (single frequency) at a 0 dbv point corresponds to a 4-mc peak deviation. Then by ordinary FM theory, noise at 4 mc at 0 dbv is -137 dbm/cbw, per noise sideband, or -134 dbm/cbw for two noncoherent noise sidebands. The 133 tandem

repeaters add 21 db for a total 4000-mile repeater noise at 0 dbv point of -113 dbm/cbw at 4 mc. The noise follows the 20 db/decade FM law, and is shown as the "Radio Repeater" curve on Fig. 6.

4.2 *Noise From FM Terminals*

FM terminal noise comes primarily from frequency modulated noise present in the output of the klystrons in the FM transmitter. At frequencies above 25 kc the noise is essentially flat with frequency at a level of four db below 1 cps rms deviation per cycle of bandwidth. This is equivalent to -133 dbm/cbw at a 0 dbv point. Below 25 kc, klystron noise rises at about 10 db/decade with an estimated value of -117 dbm/cbw at 1 kc. A 4000-mile TH route is customarily assumed to include 16 pairs of terminals. Thus the "FM Terminals" curve of Fig. 6 is drawn 12 db above the values given above for a single pair of terminals.

4.3 *Noise From Microwave Carrier Supplies*

Microwave carrier supply (MCS) noise appears in the broadband channels in two ways. In one, the noise components at signal frequency (which is 74 mc from carrier supply frequency) add directly to the signal. This source is controlled by carrier supply filtering and by balance (about 20 db) in the channel modulators. The second, more serious, form of noise injection from the MCS is at the carrier supply frequency itself. This noise is equivalent to random amplitude and phase modulation of the carrier supply frequency. The modulators, as frequency shifters, are completely transparent to this noise, and the phase modulation appears as noise at the output.

Some of this noise, generated in the low frequency multiplier chain, is reduced by a narrow-band filter in the 252-mc input to the high frequency multiplier chain. In a repeater station, the low frequency noise within the pass band of the 252-mc filter is further reduced by cancellation between the receiver and transmitter modulators. An additional contribution is from the TWT amplifier used to amplify the carrier supply for the transmitter modulator.

From laboratory measurement, the MCS noise in a repeater station is approximately uniform with frequency, at -145 dbm/cbw at a 0 dbv point. It is assumed that there are 116 such repeaters in 4000 miles. For terminals, and other special locations where cancellation does not occur, the noise rises below 1 mc to a value of -138 dbm/cbw at 200 kc and flattens off there. It is assumed that there are 16 such stations in 4000 miles. Combining the two types yields -124 dbm/cbw

above 1 mc and -122 dbm/cbw below 0.2 mc as the MCS contribution in 4000 miles, as shown in Fig. 6.

Noise contributions from the FM receiver and ion oscillations in the TWT amplifier have been neglected in Fig. 6. The contribution of 16 FM receivers is believed to be about -136 dbm/cbw, flat across baseband. Ion oscillations, which can occur in the region of 3 mc, may be extremely bad, but normally occur only immediately after a new TWT is installed and vanish rapidly as the TWT ages.

4.4 *Pre-emphasis*

It is clear from Fig. 6 that if all message channels are transmitted at the same transmission level (TL) at a 0 dbv point, the noise in the channels at the top edge of MG 3 (8.3 mc) is 12 db greater than that in channels at the bottom of MG 1 (0.3 mc). The object of pre-emphasis is to equalize this difference. Usually the contribution from second-order modulation products is taken into account also.

The TH system uses step-type pre-emphasis in which the various master groups of the telephone multiplex signal are transmitted at different levels. This avoids the need for careful match of pre-emphasis and de-emphasis networks, since the pre-emphasis is obtained by a hybrid tree in the multiplex terminal and a match can be attained by adjustment of flat gain controls. A limit on the amount of pre-emphasis is set by the nonlinearity of the baseband amplifiers, whereby second-order difference products between the higher level components fall back on the lower level part of the band.

If the pre-emphasis steps are chosen to make fluctuation noise about the same in the top channels of each master group, these steps would be 4.5 and 7.5 db. As a result of minor practical considerations, the steps used are actually 4.7 and 7.2 db.

The "load capacity" of an amplifier handling multichannel carrier telephone signals has been defined by Holbrook and Dixon[9] as the power in dbm at 0-db transmission level (TL) of the single frequency sine wave whose peaks are exceeded by the complex modulating signal only a very small percentage of the time. The same load capacity is customarily used to relate the nominal peak deviation of an FM signal to the actual complex modulating signal. Fig. 7 of Ref. 9 gives the required load capacity for various numbers of telephone channels. When extrapolated to 1860 telephone channels and modified by adding approximately a db of extra safety margin, this gives 28.8 dbm as the load capacity used in TH calculations.

Without pre-emphasis the TL at the 0 dbv point for 1860 channels would be simply -28.8 db, since a sine wave of 0 dbm at that point causes rated peak frequency deviation. But with step pre-emphasis, the TL's of the three mastergroups are different. For the same total power at the 0 dbv point, the pre-emphasized TL's are: -33.6 db TL for MG 1, -28.9 db TL for MG 2 and -26.4 db TL for MG 3.

The noise at the top of MG 1 (3084 kc) is shown in Fig. 6 as -114 dbm/cbw or -79.2 dbm in a 3-kc band. Since -82 dbm of fluctuation noise in a 3-kc band is 0 dba, MG 1 noise is $+2.8$ dba at -33.6 db TL or $+36.4$ dba at 0-db TL. This assumes implicitly that a voice channel travels the full 4000 miles on the same baseband frequency assignment. Since this is unrealistic, a more appropriate noise value is given by the rms noise across the whole baseband. This corresponds to the result that would be obtained by extensive frogging of voice channels. The rms noise is about 1.6 db less than the maximum value, or $+34.8$ dba at 0-db TL. Some noise margin is thus available for other sources between this value and the over-all objective of $+39$ dba at 0-db TL.

4.5 *Modulation Noise*

Modulation noise in the TH system is expected to be chiefly second order, arising from residual unequalized linear envelope delay distortion (EDD). The modulation produced in an FM system by linear EDD of 1 mμs/mc is:

$$M_d = \frac{10^{-15}}{2\pi} MM'$$

in which M is the baseband modulation function and M' is the time derivative of M. It is easily shown that M_d contains second harmonic terms and sum and difference frequency terms of all the modulating frequencies.

In the conventional simulation of a telephone system by noise loading, M is considered to be a large number of incommensurate frequencies of random phase, one for each unit of baseband width. The total modulation power M_d, at any given frequency is obtained by summing the powers of the individual modulation products arising from all possible combinations of fundamentals which give that frequency.

At present, in addition to a basic equalizer for each repeater section, blocks of delay slope equalization of $+1$ mμs/mc or -1 mμs/mc are provided for distribution along each switching section, as determined by over-all EDD measurements. Mop-up equalization will be added at the end of each switching section to permit over-all delay slope equali-

zation to $\pm\frac{1}{8}$ mμs/mc. It is expected that, with final equalization, variations with time will be such that on the average the linear EDD of a switching section will be no greater than $\frac{1}{4}$ mμs/mc. Assuming 25 switching sections in 4000 miles, the computed 4000-mile modulation noise is shown on Fig. 7 together with the sum of the fluctuation noises from Fig. 6. This, however, neglects other factors such as intermodulation from other sources: for example, waveguide echoes and second-order sideband clipping because of restricted bandwidth (this latter is essentially a third-order modulation effect). The circles and triangles on Fig. 7 show an extrapolation of measurements (in 1960) on the initial installation. Noise loading simulating 1860 channels, pre-emphasized, was used. Fig. 8 shows the over-all EDD of 60 hops in tandem (1900 miles) obtained by looping two two-way channels. The residual EDD slope of this characteristic is the accumulation from the eight switching sections, which are individually equalized to $\pm\frac{1}{2}$ mμs/mc only. The coarse structure is the systematic deviation of the basic equalizer from the radio repeater and is about $\pm\ \frac{1}{2}$ mμs/repeater. The

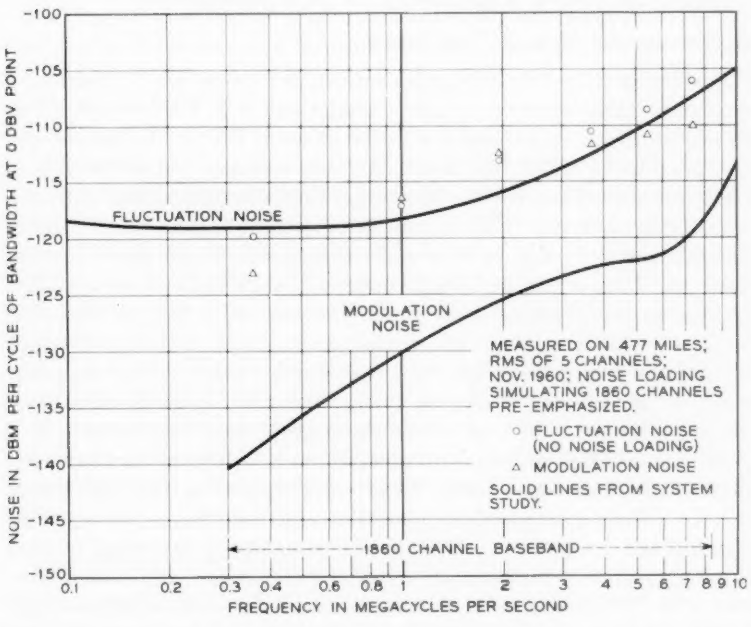

Fig. 7 — Modulation and fluctuation noise for 4000 miles of TH radio, as calculated in system study and as extrapolated from field measurements.

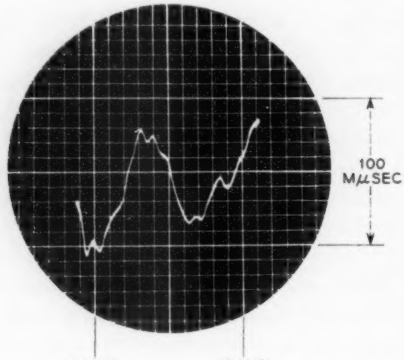

FIG. 8 — Typical over-all envelope delay distortion of 1900 miles (60 repeater sections in tandem) of TH radio.

fine structure (small ripples) is probably due to echoes in waveguide or IF cabling. The continuing work on equalization is expected eventually to reduce the EDD and the modulation noise substantially.

4.6 *Limiters and Mop-up Equalization*

Limiters are used in each radio transmitter to reduce conversion of amplitude modulation to phase modulation. AM/PM conversion is defined as 20 log (x/a), where x is the index of PM at the output and a is the index of AM at the input. Without a limiter, the conversion in a TH transmitter is around −10 db; the limiter improves this to around −30 db. In itself, AM/PM conversion is basically just another source of transmission distortion and is indistinguishable from other sources. However, when produced by active devices, such as limiters and TWT amplifiers, it is variable with time. The success of any nondynamic equalization plan requires that deviations stay put. Although dynamic equalization (regulation) has been considered, it has not yet been the object of development effort for microwave radio systems. Hence the use of the limiter in TH, to reduce the magnitude of this variable.

The ideal in any system is to equalize each individual repeater over the needed transmission band. Thus, each repeater in the TH system includes a basic gain and delay distortion equalizer. Since this equalization can not be perfect, a variable IF equalizer is provided in each switching section to mop-up the remaining distortions. Previous experience with TD-2 showed that many factors reduce the effectiveness of mop-up equalization, a principal one being the inherent instability of the system. Considerable effort has been taken in all parts of the TH system to improve transmission stability.

The use of a limiter in each station tends to "freeze-in" the distortion occurring in that repeater section. Since it is easy to postulate exaggerated examples which demonstrate that mop-up equalization through tandem limiters is impossible, extensive calculations were made to determine feasibility. These calculations, using digital computer techniques, showed that substantial benefits are potentially realizable from mop-up equalization when nominal distortions are considered.

Envelope delay distortion can be measured through limiters by the two-frequency sweep method.[10] Thus mop-up equalization to minimize the over-all EDD would apply the exact inverse of the EDD of the system. With individual repeater characteristics representable by a simple cubic power series, use of the inverse EDD correction at the end of ten sections gave a computed improvement of the order of 15 db in cross-modulation noise. The EDD correction increases the video gain roll-off, an effect typical of systems with limiters. This may be corrected with a symmetrical (with respect to the carrier) gain equalizer. The computations showed it to be essential to have a limiter between the delay mop-up and the gain mop-up equalization.

The calculations further indicated that mop-up equalization is ineffective for ripples in the transmission characteristic, i.e., for echoes. Accordingly, no provision is made for mop-up ripple equalization. Instead, one of the design objectives of TH is to keep all echoes less than −60 db, and hence not require equalization. This 60-db echo requirement is obtained from the Bennett, Curtis, Rice paper.[11] Using the TH constants of 0.7-mc mean rms deviation, 10-mc top baseband frequency, 3-db improvement from pre-emphasis and 4000 random echoes, the effect of 60-db echoes can be determined from Fig. 5.7 of Ref. 11 as 28 dba at 0-db TL, i.e., about 9 db below fluctuation noise. All return losses in the TH system have design limits of 30 db to maintain the 60-db echo limit.

It is not yet known how much of the potential benefit of mop-up equalization shown by the calculations can be realized in practice. Tests of mop-up equalization made in 1960 on the initial installation were inconclusive, possibly because of the presence of a large delay distortion, as shown on Fig. 8, which was not equalizable by a cubic power series.

V. INTERFERENCE BETWEEN RADIO CHANNELS

The design approach for TH has been to make this general source of noise small compared to fluctuation and modulation noise. With twenty radio channels packed tightly in the allocated frequency band, there are numerous possibilities for interchannel interferences. Some of these

are well known from experience with the earlier TD-2 microwave system.[12] Others are peculiar to TH.

5.1 *Classifications*

In the study of interchannel interferences, it is useful to distinguish between interference produced by the ordinary crossmodulation effect in transmitter, receiver and shift modulators, and interference inherent in limiter action. True, this is in reality a form of intermodulation due to the nonlinear characteristic of a limiter, but it is distinct from the first in that the performance of a limiter is predictable (ideal limiters are assumed) but that of modulators has to be determined by measurements.

The expression for limiter interference* has been derived repeatedly.[13] The interference at baseband due to the adjacent channel is a tone of the frequency of the channel spacing (29.65 mc), frequency modulated with the baseband signals of both channels. The tone itself falls outside baseband, but the spectrum of the interference covers a wide range and part of it falls within baseband, causing interference which is treated as noise. A consequence of limiter action is that if the input to the limiter consists of the carrier and a relatively weak interference tone, say x dbC (db relative to the carrier), on one side of the carrier, then in the output spectrum there are frequencies on both sides of the carrier, at $(x - 6)$ dbC.

According to the character of the interference as observed at baseband, interferences fall into three classes: tone, noise, and intelligible crosstalk. Tone interference rises from combinations of the various fixed frequencies in the system. The noise-like interferences arise from combinations of various signal sideband components with each other and with the carriers. The combinations are usually quite involved, and there is considerable uncertainty regarding the soundness of the computation techniques for long systems. Most measurements of this class of interference are unsatisfactory and the experimental difficulties are large. The third class (intelligible or clear crosstalk) is given the special name of "direct adjacent channel interference," and has recently been given much attention.[14]

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* The phase distortion in the wanted channel is

$$u_d = r \sin (\omega_D t - u_t + v_t)$$

where r = carrier ratio ($r \ll 1$),
ω_D = difference between carrier frequencies,
u_t = phase modulation of wanted carrier, and
v_t = phase modulation of unwanted carrier.

5.2 *Objectives*

Objectives for interferences depend on their class: tone, noise, or clear crosstalk. The allocations made in the earliest engineering studies were, for 4000 miles,

Co-channel	24 dba at 0 db TL
Adjacent Channel	21 dba at 0 db TL

No specific allocations for other types of interchannel interference were made. The general policy has been adopted of applying a 21-dba objective to new noise sources as they are found.

The objective for tone interference used in systems analysis and design is shown on Fig. 9. In the derivation, which is too lengthy to give here, −70 dbm at 0-db TL has been used as the basic requirement on single frequency interference into a voice channel. With minor exceptions, no pure tones have been observed in the 0- to 10-mc baseband due to interference from other channels or to spurious frequencies on the microwave carrier supply. This is because of the inherent instability (jitter) of an FM carrier produced by beating two 6-kmc klystrons. A steady FM carrier with a steady interference tone results in a steady baseband tone. When the FM carrier has jitter, the energy at baseband, instead of being concentrated at a single frequency, is smeared over a frequency band. Rather than the tone requirement, then, a 10-db more lenient noise requirement may be used. The noise, as heard on telephones when demodulated to voice, has an unpleasantly harsh

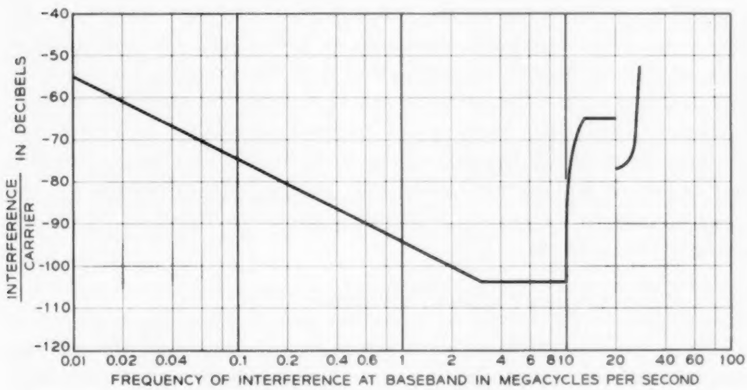

Fig. 9 — Objective for single-tone interference at IF at limiter input.

and burbling quality, completely different from the steady hiss of thermal noise.

For clear, intelligible crosstalk, as produced by direct adjacent channel interference, the objective is quite severe. Its derivation, due to H. E. Curtis, is given in the Appendix. In terms of the coupling loss per repeater, as measured at baseband, the objective is 94-db average during free space transmission, and 40-db for the minimum during a 27-db fade. Laboratory data on the TH system and field data on the TD-2 system indicate that the coupling increases (loss goes down) two db for one db of fading. Therefore, the second objective also corresponds to 94-db minimum loss under free space conditions.

5.3 *Laboratory and Field Observations*

The results given here are a combination of calculations, laboratory measurements of single exposures, and data obtained on the initial installation of TH between Prospect Valley, Colorado, and Salt Lake Junction, Utah, in 1960. The classification below is based primarily on the source of interference and is arranged roughly according to the separation between the wanted and unwanted channels.

5.3.1 *Co-channel Interference*

This refers to interference from other repeater sections using the same carrier frequency. Basically the only protection against this is antenna directivity and physical separation. The former has its limitations. In particular the inherent directivity of an antenna can be, and is, spoiled by reflections of the transmitted microwave energy by objects (e.g., large buildings, mountain ridges) along the transmission path, usually in the general neighborhood of the antenna. The front-to-back ratio of horn-reflector antennas at 6 kmc will therefore show a statistical distribution which is not presently available. However, calculations show that if the rms of this is 65 db, the co-channel interference objective will be met with 2-db margin. Measurements on the initial installation of twenty F/B ratios gave two values of 61 db, all others 66 db to 85 db.

5.3.2 *Direct Adjacent Channel Interference*

This is interference in which a signal on the unwanted channel appears unchanged on the observer's channel. Extensive laboratory tests of direct adjacent channel interference (DACI), made on a single radio repeater, result in the following views on DACI production:

i. The adjacent channel carrier and its sidebands produce in the limiter sidebands around 74 mc which are roughly half AM and half PM. The limiter is the chief source of DACI.

ii. The limiter has a "DACI constant," analogous to a modulation coefficient, which permits the calculation of the PM component of the interference sidebands near the carrier and at the limiter output from the known adjacent channel interference spectrum at the limiter input. However, the physical mechanism of the sideband production is not yet understood.

iii. Any amplifier in compression acts like a poor limiter and can contribute to DACI production.

iv. DACI increases 2 db (very closely) for 1-db increase in carrier ratio. Carrier ratio is defined as the level in db of the unmodulated carrier of the unwanted channel with respect to that of the wanted channel at the input to the latter's receiving channel separation network. An increase in carrier ratio corresponds to an increase in unwanted signal.

v. The interference is affected by the selectivity of the radio receiver, determined in part by microwave filters and in part by the IF amplifiers. Thus DACI is worst at high baseband frequencies.

vi. The laboratory measurement gives 40 db as the equal-level coupling-ratio per radio receiver at 8 mc for 0-db carrier ratio. With 25-db cross-polarization discrimination, the per exposure coupling is 90 db, or quite close to the objective derived in the Appendix.

Measurements made on the initial installation at −3-db carrier ratio on five repeater sections showed considerable variation in equal-level coupling ratios, ranging from 40 db to 58 db. On the whole it appeared that the laboratory value of 40 db at 0-db carrier ratio at 8 mc is representative. It was quite definitely established that voltage addition in tandem sections [see Appendix, 1e] does not hold, in which case the controlling requirement is probably the 40-db value during a 27-db fade [Appendix, 2d]. With 40 db at 0-db carrier ratio, 27 db of cross-polarization discrimination is needed, a value generally obtained on this installation. Fig. 10 shows the distribution of cross-polarization discrimination.

A few measurements were made with normal couplings over the whole route (15 hops). Here it was necessary to use a very strong drive to find the interference in the noise. Depending on the selection of unwanted and wanted channels, the over-all 8-mc coupling varied between 73 db and 95 db. Even the 73-db value meets the free-space transmission requirement of the Appendix.

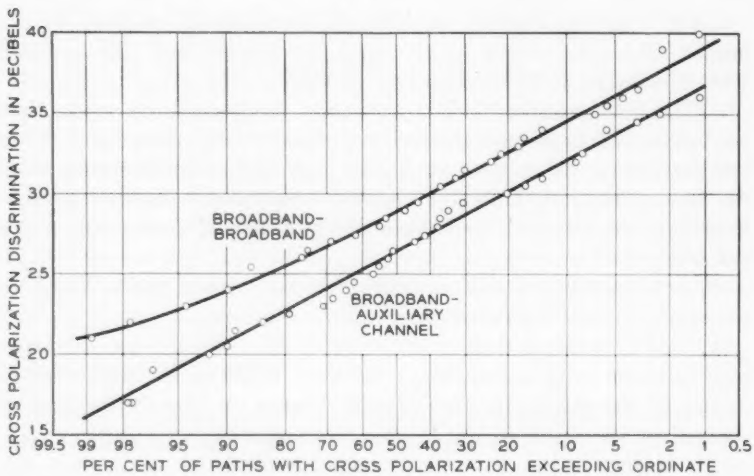

Fig. 10 — Distribution of cross-polarization discrimination, as measured on the initial installation.

5.3.3 *Adjacent Channel Interference*

This type includes the effects of (1) cross-modulation in the receiver modulator and (2) the inherent cross-modulation produced by ideal limiting. In the first case the chief possibility for interference arises from intermodulation between the carrier spikes, which produces "tones" at low baseband frequencies (typically 100 kc or less) or around 15 mc. Cross-modulation of the sideband energy is negligible. On the basis of the single-tone requirement and using the measured modulation performance of a very early model of the TH receiver modulator, requirements were calculated on unwanted tones at the modulators, and these were used to establish requirements on filters in the microwave carrier supply and in the signal path.

With regard to limiter action, elaborate computations on this indicated that the interference would meet the allocation of 21 dba with 20 db of cross polarization discrimination, provided the top baseband frequency was less than $\frac{1}{3}$ channel spacing, or 9.88 mc. This condition is satisfied in TH. The chief components of interference are, in order, due to (a) each adjacent carrier with first-order sidebands of the other adjacent channel, (b) second-order sidebands of adjacent channels with first-order sidebands of the center channel and (c) direct interference from third-order sidebands of adjacent channels. The "tone" due to

interaction of the two adjacent carriers was dismissed as being below the lowest telephone channel. The chief uncertainty about the calculations is with respect to the addition of interferences in tandem sections; the calculations assume power addition.

Both laboratory and field measurements of this type were unsatisfactory because of measurement difficulties. Such data as were taken indicate that it is extremely unlikely to be an appreciable source of interference.

5.3.4 *Lost Carrier*

If the unwanted channel loses its carrier, the TWT amplifier saturates at full power with noise spread over a wide band (normally about 98 per cent of the total radiated power is concentrated in the carrier). In the adjacent channel, 30 mc away, this noise power is very much greater than the sideband energy usually present, and if no action were taken would make the channel uncommercial. Protection against this is given by providing an automatic carrier resupply circuit in each radio transmitter.

5.3.5 *Tertiary Interference* (*Limiter transfer*)

The basic mechanism of this is shown in Fig. 11. The limiter in channel 27 transmitter transfers the interference due to channel 18 to the other side of the channel 27 carrier, producing interference into channel 26 receiver. By this mechanism, a 5-mc baseband tone on channel 18 becomes a 24.65-mc interference on channel 27, which in turn becomes a 5-mc interference on channel 26. The interference frequency on channel 26 differs from the original frequency on channel 18 by only a small amount, due to variations of the actual carrier frequencies of channels 18, 17, 27 and 26 from their theoretical values.

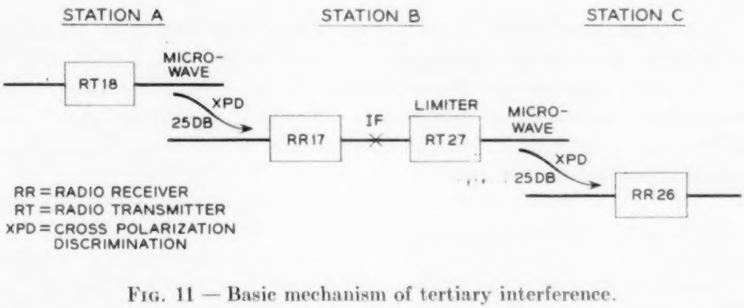

Fig. 11 — Basic mechanism of tertiary interference.

Laboratory tests show that limiter transfer action is perfectly straightforward and in exact accordance with elementary FM theory. The seriousness of limiter transfer action was first noted in dealing with interference from the auxiliary channels into broadband channels. Eventually the antenna connections of the auxiliary channels were changed to minimize the tertiary interference from them; see Section 5.3.8, below.

In a slightly different type of tertiary interference, the auxiliary channel carrier is picked up by the broadband receiver of the adjacent channel, amplified and reradiated by the broadband repeater; here the channel separation is only 20 mc. This reradiated signal is picked up by the auxiliary channel receiver in the next station. Slight differences in the frequencies of the various crystal-controlled oscillators involved cause interference between the carriers to appear as a tone which may fall into the band assigned to TH switching signals. The interference is normally not objectionable, but during a fade of the broadband channel (or worse yet, when the preceding broadband transmitter is shut down for maintenance), the AGC of the broadband receiver brings up the spurious signal to the point where the interference is comparable in level to the switching tones. This situation is treated by the insertion of a narrow-band cavity resonator absorption trap at 94 mc in broadband channels 1 and 8, and by holding the frequency tolerance of the crystal oscillators to values which cause the interference to fall at a frequency below that of the switching tones.

The chief protection against interferences of types 2 to 5 is cross polarization discrimination, and it is thus important to avoid its deterioration. Some discrimination is realized from the selectivity characteristics of the channels.

5.3.6 *Image Interference*

The image suppression given by microwave filter selectivity (channel separation plus channel bandpass) is about 100 db, and in addition 25 db of cross polarization discrimination is effective. This much discrimination makes image interference completely negligible, since it is of the same nature as co-channel interference for which 65-db coupling loss is considered acceptable.

5.3.7 *Carrier Supply Interferences*

Interferences which are basically of the tone type are produced by interactions among the various high level carriers used to drive the

various modulators. The most difficult problems here were solved by the adoption of the coherent carrier supply. Due to slight variations of the actual frequencies from their theoretical values, both as generated in FM transmitters and as modified by repeated frequency shifts up and down in successive repeater sections, the interferences from various sources will be spread out over relatively narrow bands, at most a few hundred kc wide, centered on the carrier and on 14.82 mc. When the TH system is operating properly, the interferences near the carrier and the second-order difference products between those at 14.82 mc will all fall below the lowest telephone channel at 300 kc.

Interference arises from carrier leak through the transmitter shift modulators on channels 1 and 3, which results in those channels transmitting also on channel 2; channels 6 and 8 similarly interfere into 7. The channel 2 spectrum is badly distorted by the channel filters, but with high modulating frequencies, one set of sidebands gets through fairly well and modulates with the regular channel 2 carrier to form the interference. Field measurements indicate that this interference will be about 11 dba at 0-db TL, provided 20-db or more carrier balance is maintained in the shift modulators.

In the field excessive interference at 59.3 mc was observed in channel 8; this turned out to be direct radiation at 6 kmc from the standby microwave carrier supply into the channel 8 receiver modulator; these are in adjacent bays, and the radiation path bypasses the carrier supply filter. This is remedied by proper shielding.

Spurious tones at 420 kc and 5.08 mc were observed on all channels at baseband. The source of these was eventually located in occasional units of the 14.8-mc crystal oscillator in the microwave carrier supply. The tones occur as the result of a beat between the third and fifth mechanical overtones of the crystal, which frequency modulates the normal output. Slight modification of the circuit remedied this.

These last two interferences are classic examples of the need for extensive testing on a commercial installation, since neither was anticipated in systems design or found in laboratory testing.

5.3.8 *Auxiliary Channels*

These channels were a troublesome source of interference because their frequency assignments are not harmonics of 14.82 mc. Interference into the broadband channel is considered under control except possibly where bad building reflections reduce antenna side-to-side coupling loss. The controlling interference here is tertiary, from channel 20 transmitter into channel 17 receiver, via channel 18, producing 4.87

mc in channel 17. To meet the interference requirement, 90 db separation is required between channel 20 transmitter and channel 18 receiver. The only way to meet this is to put these channels on separate antennas; furthermore, channel 20 has to be cross-polarized with respect to channel 21; this establishes the channel assignments of Fig. 4. Fig. 12 shows a distribution curve of antenna couplings.

The chief potential source of interference into auxiliary channels from the broadband channel is sideband energy in the adjacent broadband channel. Under normal conditions, the cross polarization discrimination puts this sideband energy at least 10 db below fluctuation noise level.

5.3.9 *Envelope Delay Sweep*

This test signal, used for the measurement of envelope delay distortion at IF, puts large amounts of signal energy at the band edges, and experience on the initial installation shows it to be a potential source of severe interference. For the purposes of collecting data, a sweep of ± 14 mc (60 mc to 88 mc) was used, but this creates intolerable conditions on the auxiliary channel. To avoid trouble with an operating system, the sweep has to be limited to ± 10 mc.

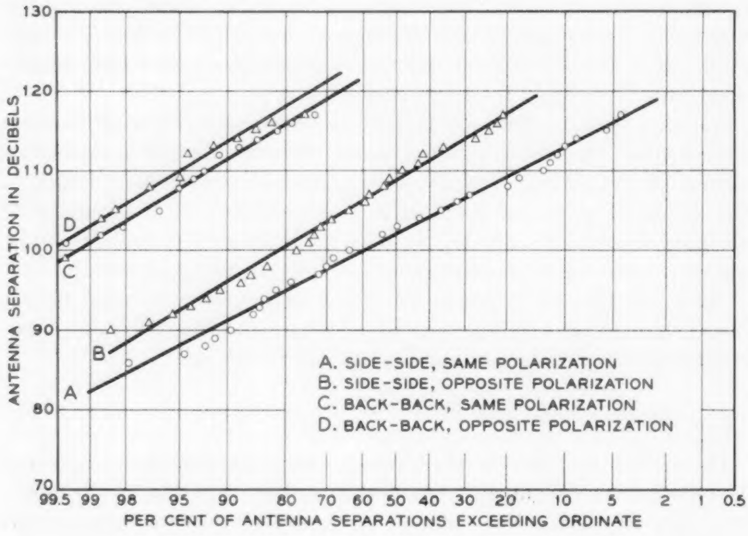

Fig. 12 — Distribution of antenna coupling loss, as measured on initial installation.

5.4 *Conclusion*

The initial installation revealed a few unanticipated sources of interchannel interference. When remedial measures for these have been installed, interchannel interference will not normally be a problem in TH with careful maintenance, and with antenna separations as good as obtained on the initial installation.

VI. ACKNOWLEDGMENTS

The development of a system as massive as TH, with its complex of subsystems, requires the contributions of literally hundreds of individuals, creating, adapting, building and polishing both electrical and mechanical designs. The authors of this paper, and of the other papers in this issue, as reporters of the results, acknowledge these contributions without attempting to list every individual by name.

APPENDIX

*Objective for Direct Adjacent Channel Interference**

During periods of free-space transmission, all repeaters contribute more or less equal amounts of crosstalk to the over-all system. On the contrary, during fading periods it is uncommon for more than one repeater at a time to fade 30 db or more. During such a period, then, nearly all the crosstalk will come from one repeater. Consequently it is reasonable to set up two objectives, viz., one for free space transmission and another for a 27-db fade (the point at which the system switches to the protection channel).

Telephone Crosstalk

1. *With Free-Space Transmission*

a. System noise − 4000 miles, 0 db TL = 39 dba

b. Assume that crosstalk from a "1 per cent talker" must be 10 db below the noise, or = 29 dba

c. A 1 per cent talker reads at 0 db TL = 86 dba

d. Coupling at baseband (4000 miles) = 86 − 29 = 57 db

e. Assume voltage addition for each repeater contribution up to 1000 miles and power addition for each 1000 mile portion; correction to a "per repeater basis" is then = 37 db

* Original derivation is due to H. E. Curtis.

f. Allowable average coupling per exposure, as measured at baseband = 94 db

2. *During a 27-db Fade.*

Assume that the controlling amount of crosstalk comes from the faded section. It would be extremely rare to have a 1 per cent talker and a 27-db fade simultaneously; therefore it will be assumed that only the "rms" talker must be maintained well below the noise. Proceeding as before,

a. Noise due to one section during a 27-db fade = 44 dba

b. Assuming the "rms" talker must be 8 db below noise, or = 36 dba

c. An "rms" talker reads at 0 db TL = 76 dba

d. Allowable maximum coupling per exposure for a 27-db fade, as measured at baseband = 76 − 36 = 40 db

Television Crosstalk

The 4-mc coupling loss at which 50 per cent of the observers in subjective tests rated television crosstalk "just perceptible", was 58 db for flat coupling and 40 db for coupling with 3 db per octave slope. Since the latter corresponds more nearly to the observed coupling, the TV requirement is less severe and is not controlling.

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TH Radio System Equipment Aspects

By P. T. HAURY and W. O. FULLERTON

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This article discusses the considerations governing the general over-all physical arrangements of the TH system, which lead to differing treatments for the radio equipment, FM terminals, and protection switching equipment. It describes the general mechanical features of these divisions, from the sliding-rack bay used for the radio equipment to the extensive use of plug-in units for protection switching.

I. FUNDAMENTAL EQUIPMENT PLANNING

It is the nature of systems such as TH radio that the problems associated with engineering station layouts, and with preparation of orders for production in the manufacturing shops, are of considerable magnitude. More obviously, the designer is concerned with arrangements for operation and maintenance, with manufacturing and installation problems, and with expenses involved in all of these. Reliability is also of particular importance in the TH system in view of the large message circuit capacity of a TH radio channel. The design of the TH equipment is based on satisfying the physical problems while retaining proper economic balance.

To reduce engineering and installation effort, the equipment is designed insofar as practical for assembly, wiring, and testing of complete bays in the manufacturing shop. These operations can be performed more efficiently there, and the amount of field installation and testing effort is thereby reduced. Engineering effort is also reduced by adherence to standardized floor-plan arrangements wherever possible.

For convenience and speed in the restoration of service in the event of failure, extensive use is made of plug-in connections and plug-in units. These are also used for efficient packaging and ease of installation as a system grows. Furthermore, the plug-in designs are consistent with the philosophy of repair procedures. Spare units are provided in each station or at nearby maintenance centers so that defective units can be readily replaced and sent to a maintenance center for repair. Some maintenance

centers are equipped with a test bench and facilities for testing units and making repairs.

The design of bay equipment is directed toward arrangements which provide ready access for routine testing and replacement of units. Where possible, the transmission equipment is given preferential location on the bays to minimize the use of ladders during normal maintenance. Routine tests are made with integrated sets housed in rolling cabinets and with multipurpose meters which plug into metering connectors on the bays.

To assure reliability, great care has been used in the choice of components. This is particularly true in the case of capacitors, which are hermetically sealed designs wherever possible. All connectors are provided with goldplated contacts and have been rigorously tested. Reliability is also stressed in the power plants and, as will be discussed later, in the blower system for the microwave radio bays.

In addition to the topics discussed below, a number of miscellaneous items should be mentioned for the sake of completeness, since they will not be considered in detail. They are, however, important parts of the complete system, and include alarm and order wire equipment, so vital to operation and maintenance, and patching and testing jack arrangements.

II. TH STATION ARRANGEMENTS

There are two general types of TH radio stations: (a) main radio stations and (b) radio repeater stations.

Main stations are varied in their characteristics. They may be end points or intermediate points in the system where intermediate frequency (IF) signals are accepted from and delivered to terminal facilities. In addition, main stations provide points of flexibility in the system where provision is made for IF patching and monitoring, automatic protection switching, dropping or picking up of signals to and from local or spur radio facilities, and maintenance switching from regular to protection channels. Such stations also provide convenient locations for performing systems tests. These stations usually are fully or partially attended.

The radio repeater stations provide transmission gain and maintain line-of-sight paths. They comprise the majority of the stations in any large system and are normally unattended. Unlike main stations the repeater stations are more uniform in their makeup and are more adaptable to standardized floor plans and building construction.

The transmission equipment for the TH radio system falls into two somewhat different categories. One includes the microwave radio and

associated microwave carrier supply equipment, which is common to all stations. A general view of this equipment is shown in Fig. 1. The other includes facilities such as protection switching and patching bays which are found in main stations only. The two categories are located in different but preferably adjacent areas in main stations, as indicated by the

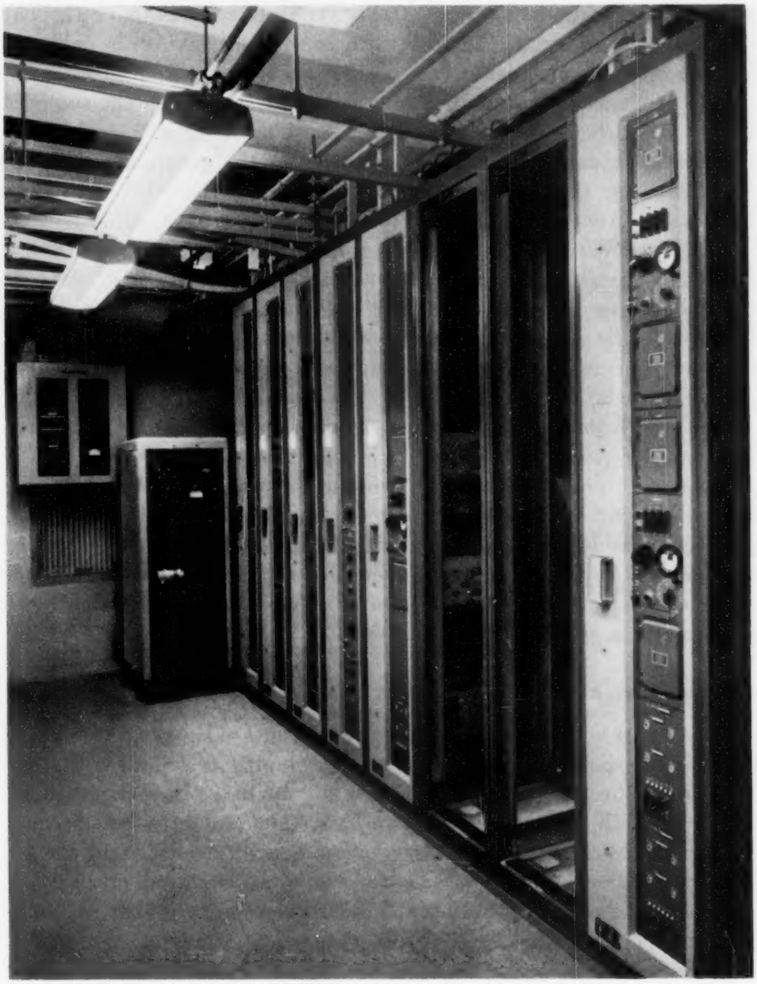

Fig. 1 — General view of the microwave radio equipment bays in a station lineup.

typical floor plan in Fig. 2. This is desirable because the requirements for the microwave equipment are determined by the number of radio channels provided, whereas the requirements for the other equipment are determined by the manner in which the radio channels are assigned and administered. Furthermore, there are unique requirements of the radio equipment which make standardization of the layouts desirable if not essential. Standardization of this type allows growth of the bay equipment in an orderly manner, with consequent savings in engineering and installation expense.

The plan adopted for the microwave bays is essentially the same for all stations, whether main or repeater. As shown on Fig. 2, each of the two lines originates with basic equipment for the group, one with the three bays for the microwave carrier supply and the other with three for the auxiliary radio channels. Beyond these basic bays, either line can grow independently as long as even-numbered channels are installed first and then odd-numbered channels. This is possible because the bays beyond the first three are associated with one side of the station for one line-up and the opposite side for the other line-up. Thus, orderly growth of a line-up is maintained even though the station may be temporarily a terminal point on the route.

The additional facilities provided at main stations vary to suit differing requirements for service and administration. These consist of FM terminals, automatic protection switching, equalization, and patching and monitoring. With the varying amounts and types of such equipment, practical considerations do not permit a high degree of standardization of floor-plan layouts. However, there are restrictions on transmission cable lengths and a preferential association of certain bays, which result in a limited amount of standardization.

2.1 *Microwave Equipment Floor Plan*

The factors which lead to standardized layouts of the radio equipment are the waveguide for microwave carrier supply distribution and for antenna connections, the tube-cooling air facilities, and the need for orderly growth. Of these factors, the waveguide facilities benefit most from standardization. Rigid waveguide is a difficult medium for ordinary field installation since the various pieces must be fitted precisely in place. Pre-engineering of waveguide parts is necessary if they are to be delivered on the job ready for installation.

The antenna waveguide runs, which originate at the broadband radio bays, are simple and direct in themselves; there are four waveguide

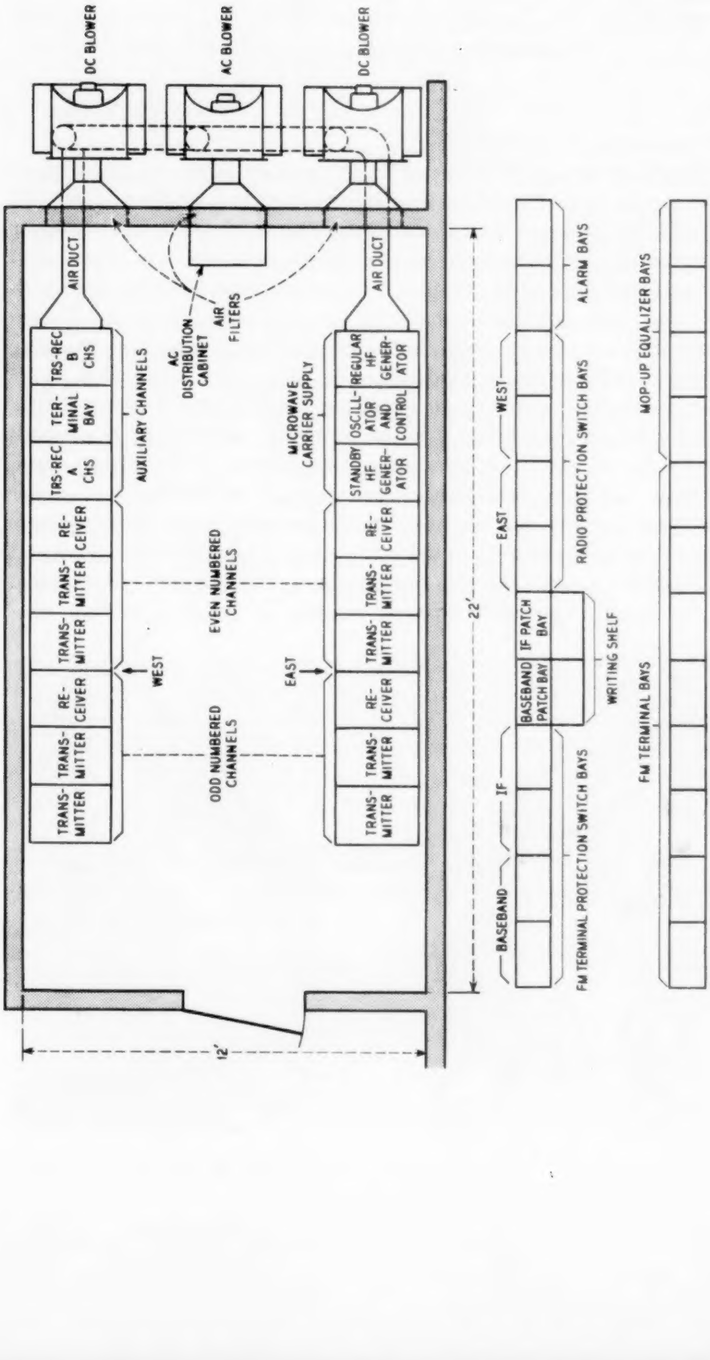

Fig. 2 — Typical floor plan of a main station, showing locations for two categories of equipment.

runs from each row of bays for a fully equipped station. However, these runs include the channel separation networks for the auxiliary channel transmitters and receivers. Careful planning in the location of the networks and in the running of the waveguide from the networks to the auxiliary channel radio bays is necessary to keep waveguide runs as short and simple as possible.

In the distribution of the carrier frequencies from the microwave carrier supply bays to the various transmitter and receiver bays, still more difficult problems arise. One of the two microwave carriers must be distributed to all the broadband radio transmitter bays (eight ultimate) and the other to all the broadband radio receiver bays (four ultimate), and both carriers to the auxiliary channel radio bays. This is accomplished by two "waveguide trees," one for each carrier. Utilizing waveguide junctions having one input and two output ports, plus waveguide spacers and bends, a single output at the microwave carrier supply bay is divided and subdivided. In the ultimate, twelve outputs are produced for the various bays containing transmitters and eight for those containing receivers. Still further splitting of the microwave carrier power takes place within the broadband transmitter and receiver bays.

The waveguide runs are located in the space above the radio bays and over the aisle between the two rows of bays. Fig. 3 shows the arrangement for a fully equipped two-way station. It is apparent that con-

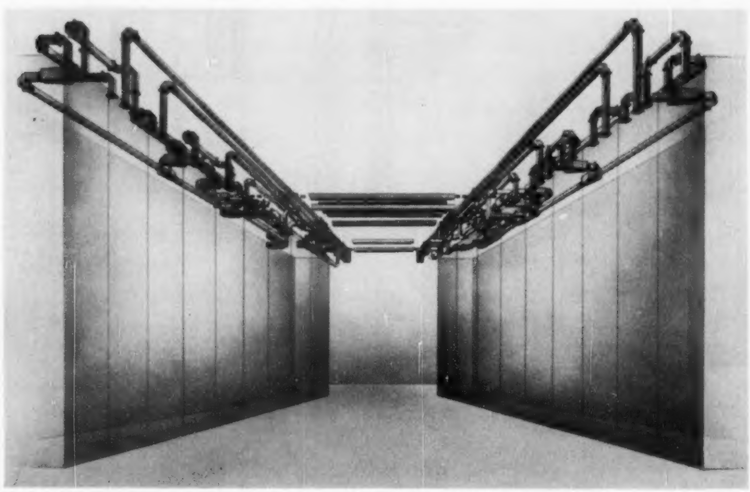

Fig. 3 — Waveguide connections for a fully equipped two-way station.

siderable reduction in engineering effort is realized if these waveguide arrangements are standardized for all stations. This has been done and standard floor plans have been prepared accordingly.

2.2 *Antenna Waveguide Connections*

Standardization of waveguide arrangements has been confined to the radio repeater room where control is practical. Beyond the radio room, the antenna waveguide runs cannot be confined to a fixed pattern because of varying job conditions. Specifically, the location of the weather seal, where the waveguide passes through the outer wall of the building, is determined by the position of the building relative to the antenna tower, and is chosen to provide the most direct run to the tower. Waveguide runs from the radio room to the weather plate must therefore be engineered to fit local conditions. This aspect of the job has been simplified by the introduction of flexible waveguide at the weather plate. The use of flexible waveguide at this point makes the mating of indoor waveguide runs to outdoor waveguide runs easier mechanically. In addition, the flexible waveguides are long enough so that any indoor waveguide in a group of eight may be connected to any outdoor waveguide in a corresponding group. This is important since the cross connections vary from station to station, as determined by many factors which are difficult to coordinate.

III. ANTENNA SYSTEM

At the weather seal in the outside wall of the radio station, the indoor waveguide runs from the radio room connect to pressure windows. At the outdoor end of the pressure windows, the pressurized waveguide runs begin; these connect to the antenna system. The horn-reflector antenna and systems combining networks of the antenna system have already been described.[1,2] Here note is made of the problems involved in the support of networks and waveguide on the towers and of waveguide runs into the buildings.

The discussion requires examination not solely of TH radio but of the total picture of TH combined with other systems. The horn-reflector antenna can transmit all three common carrier microwave systems developed for Bell System use. These are at 4 kmc (TD), 6 kmc (TH) and 11 kmc (TJ). As described in Ref. 2, systems combining networks permit coupling signals of any or all of the three systems with either horizontal or vertical polarization on a single antenna. Each such assignment requires a separate waveguide run to the pressure windows.

The number of antennas required for a given station depends on factors beyond the scope of this discussion but ranges from one to a usual maximum of eight. The usual practice is to place all antennas on a single tower, which may range in height from about 40 to 400 feet. Thus, at a station at the intersection of two routes with all three systems appearing and both polarizations used, there will be eight antennas. Each antenna will have connected to it the full array of systems combining networks, to each of which six rectangular waveguide runs will connect or an ultimate of 48 for the station. In contrast, a repeater station on a single route equipped only for TH radio need only have one combining network and two rectangular waveguide runs per antenna. The fully equipped stations with four antennas would have an ultimate of eight rectangular waveguide runs to the pressure windows.

Because of the wide variations possible in the provision of systems combining networks and the associated waveguide runs to the building, it is essential that the design of the supporting structures permit flexibility in engineering to meet local conditions. On the tower, this involves the positioning of restrainers and protective shields for the combining networks which are, in effect, an extension of the vertical run of circular waveguide from the antenna. Adjacent to each input or output derived in the systems combining networks, hangers are provided to support the rectangular waveguide extending to the base of the tower. These hangers must permit varying the position of the rectangular guide to accommodate orientations of the combining networks as determined by antenna orientation and network polarization in combination.

At the base of the tower, the rectangular waveguides are channeled into horizontal runs to the building. Where possible, the horizontally run waveguides are placed in a common supporting structure. However, the more usual condition is to provide for separate paths to the building for the different systems. This keeps the total length of waveguide required to reach the transmitters and receivers to a minimum, to achieve the lowest practical transmission loss.

IV. MICROWAVE RADIO EQUIPMENT

The microwave radio equipment, as distinguished from FM terminals and protection switching, comprises the common microwave carrier supply and the transmitters and receivers for both the broadband and auxiliary radio channels. Because of the problems involved in the maintenance of this equipment and considerations arising from use of traveling-wave tube (TWT) amplifiers in TH, a new type of bay framework has been designed. This new design is called a sliding-rack bay framework

and is also used, for uniformity, for the multiplex terminals of the auxiliary radio channel.

4.1 *Sliding-Rack Bay*

The sliding-rack bay framework permits a closely coordinated association of the radio equipment bays, uses floor space more efficiently, and provides access to all equipment from one aisle. The design also simplifies the distribution of cooling air for electron tubes and permits location of the TWT amplifiers for convenient maintenance. The design provides a conventional type of rack structure on which the equipment is mounted. This equipment rack is in turn mounted in a basic framework on rollers, with the face of the rack at right angles to the front of the bay. When access to the equipment is required, the rack is pulled forward into the aisle space. In this position both sides are fully accessible for maintenance. The bays may be arranged side-by-side in rows, like books on a shelf; and because the slideout feature permits access to all equipment from the front aisle, the rows of bays may be installed back-to-back or against a wall.

The design is illustrated in Fig. 4. The basic framework is made up of top and bottom castings which are joined together by four corner posts. The same casting is used for both the top and the bottom and is designed to result in a trough across the top of a line of bays for the running of cable. At the bottom of the bay a cover plate is provided over the trough to form a duct which runs through the line-up of bays and is extended as bays are added. This duct is used for the distribution of cooling air to the bays in the line. The sliding rack is conventional except that the rear upright is extruded with a cross section which forms a duct from top to bottom of the rack. This duct is used for the distribution of cooling air to the equipment mounted on the rack. Air is supplied to this duct through a flexible hose connected to the main duct in the base of the bay. The rack is mounted in the basic framework on heavy-duty ball-bearing tracks. Sufficient travel is provided in the tracks to permit the bay to be fully exposed in the outboard position.

The bay is $19\frac{1}{2}$ inches wide, 28 inches deep, and nine feet high over-all. All structural members of the basic framework and sliding rack are made of aluminum. The vertical members are extrusions and the top and bottom members of the basic framework, as stated earlier, are castings, also of aluminum.

To permit the sliding rack to be moved outward, the power, coaxial and miscellaneous connections are made to the rack in flexible cables which are positioned and formed to allow the required lateral movement

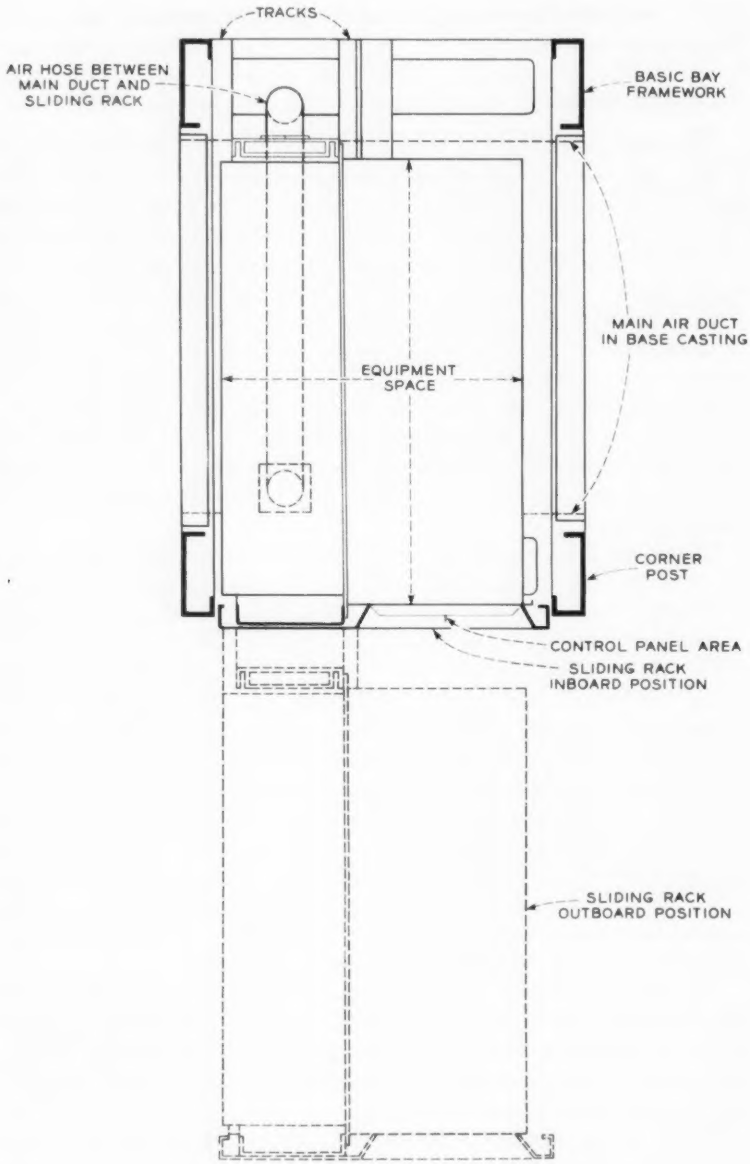

Fig. 4 — Plan view of sliding-rack bay framework.

without placing undue stress on the cable forms. Where waveguide connections are made, short pieces of flexible rectangular waveguide are provided in the waveguide runs to permit the waveguide to flex with the movement of the rack.

An area is provided on the front face of the sliding bay for operating and routine maintenance controls and for providing access to the electron tubes of the TWT amplifiers. Access to a traveling-wave tube is through a hinged door at the front end of the amplifier housing. Since the tube is approximately 12 inches long, sufficient clearance is provided in front of the door, to permit the tube to be inserted and withdrawn from the amplifier, by mounting the amplifier on the bay with the hinged access door facing the aisle. The controls and access doors are visible and accessible with the bay in the inboard position. This permits most routine maintenance checks to be made without pulling the bay to the outboard position.

Inter-unit wiring for the bay is in a local cable located at the rear upright. Connections from the sliding bay to external points, including waveguide, are attached at approximately the rear vertical center of the bay and follow an upward path (bay in closed position) to the underside of the top casting. Here, they turn forward to the front and through a slot in the top casting. All external connections terminate in this area — the waveguide in fixed flange, the coaxial cables in jacks, the power cables with sufficient length to reach the power distribution duct, and the miscellaneous signal and alarm leads on terminal strips. Thus, all installer connections are made at the top and external to the bay. No wiring or assembly functions are performed by the installer inside the bay except where equipment is added to partially equipped bays in the field.

4.2 *High-Frequency Generator Bay*

Typical of the application of the sliding-rack bay is the high-frequency generator bay shown in Fig. 5. This is one of two such bays used with an oscillator and control bay to comprise the common microwave carrier supply. The salient features of the bay design described above are illustrated in this photograph. In addition, the treatment of waveguide components and their interconnection on the mounting panel can be seen. In the upper portion of the bay are mounted panels containing the low-voltage rectifier units supplying heater voltage and the lower plate supply voltages to the electron tube circuits. At the bottom of the bay the high-voltage power supply for the TWT amplifiers appears. To protect personnel against the hazard of high voltages within this rectifier

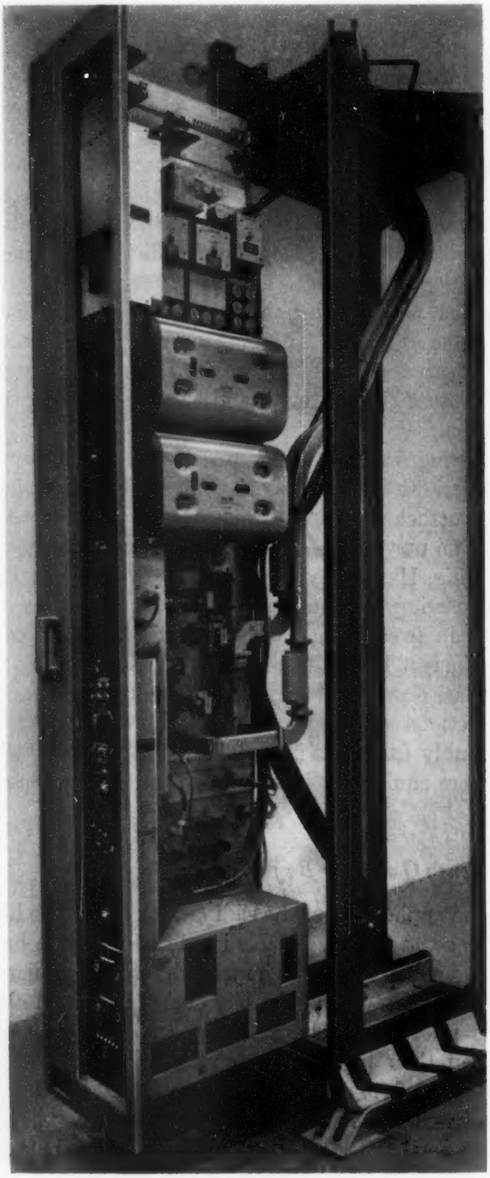

Fig. 5 — High-frequency generator bay, a typical application of the sliding-rack designs.

and the TWT amplifiers, the units are both electrically and mechanically interlocked. The electrical interlock is by means of switches released on opening of the TWT access door. The primary protection, however, is the mechanical interlock using keys which must be obtained from the high-voltage rectifier. The keys are seized in position on the rectifier until the main switch for the rectifier has been turned off.

4.3 *Radio Transmitter Bay*

Similar in general design to the high-frequency generator bay is the broadband transmitter bay illustrated by Fig. 6. Two transmitter panels and associated TWT power supplies are contained in the fully equipped bay. The low-voltage power supplies are mounted on the individual transmitter panels. Near the rear edge of the transmitter panel is seen the waveguide network (the channel separation network) which couples the transmitted microwave signal into the antenna run. Each such network is a series element of the waveguide run to the antenna. For the maximum of four broadband transmitters possible on one antenna waveguide run, two bays are needed, and the interconnection requires that the waveguide be looped through the first of the two bays to reach the second. For the fully equipped two-way station, one such pair of bays is required for the odd-numbered broadband channels and one for the even-numbered, for each direction.

4.4 *Radio Receiver Bay*

The broadband receiver bay appearing in Fig. 7 shows a somewhat different treatment of the panel design. Here the waveguide components are confined to the right side of the panel, while the left side is utilized for the IF main amplifier and equalizer. Again the channel separation network appears near the rear edge of the panel. Unlike the case of the broadband transmitter, four receivers can be mounted in the fully equipped bay, and looping of the antenna run through the bay is not necessary. Similarly, only half as many receiver bays are required for a station.

4.5 *Auxiliary Channel Radio Bay*

For the auxiliary radio channels, the transmitter and receiver for a send and receive pair to one side of a station are combined on a common panel, and two such panels are mounted in the same bay at a through station. The bay arrangement is shown in Fig. 8. Since the auxiliary channels comprise two pairs in each direction, representing four terminal

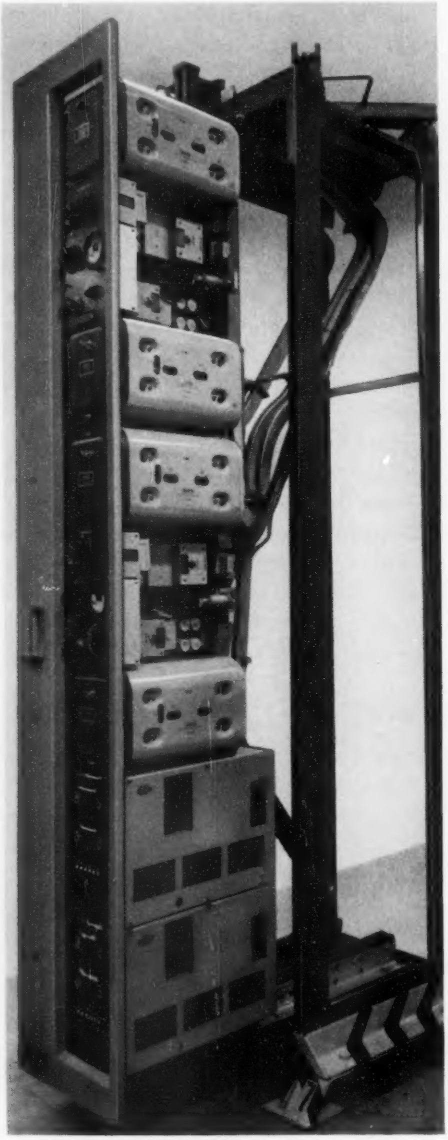

Fig. 6 — Broadband transmitter bay.

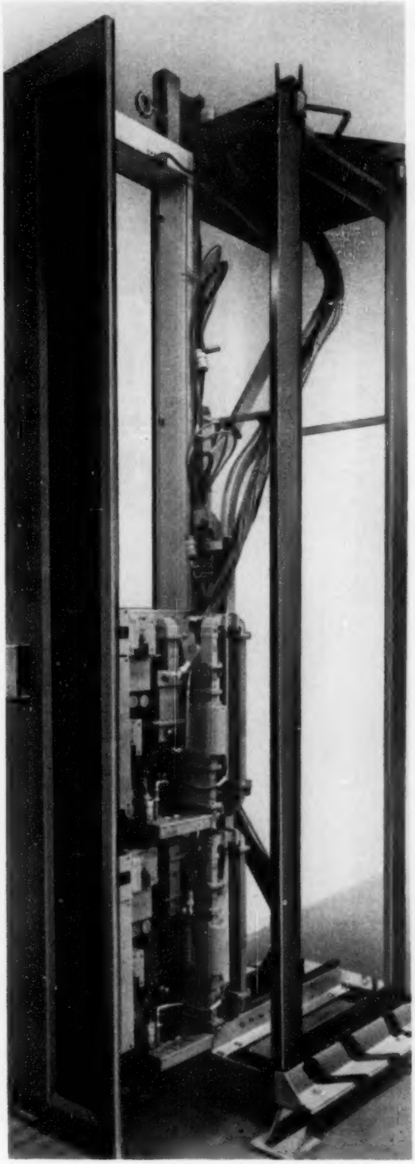

Fig. 7 — Broadband receiver bay.

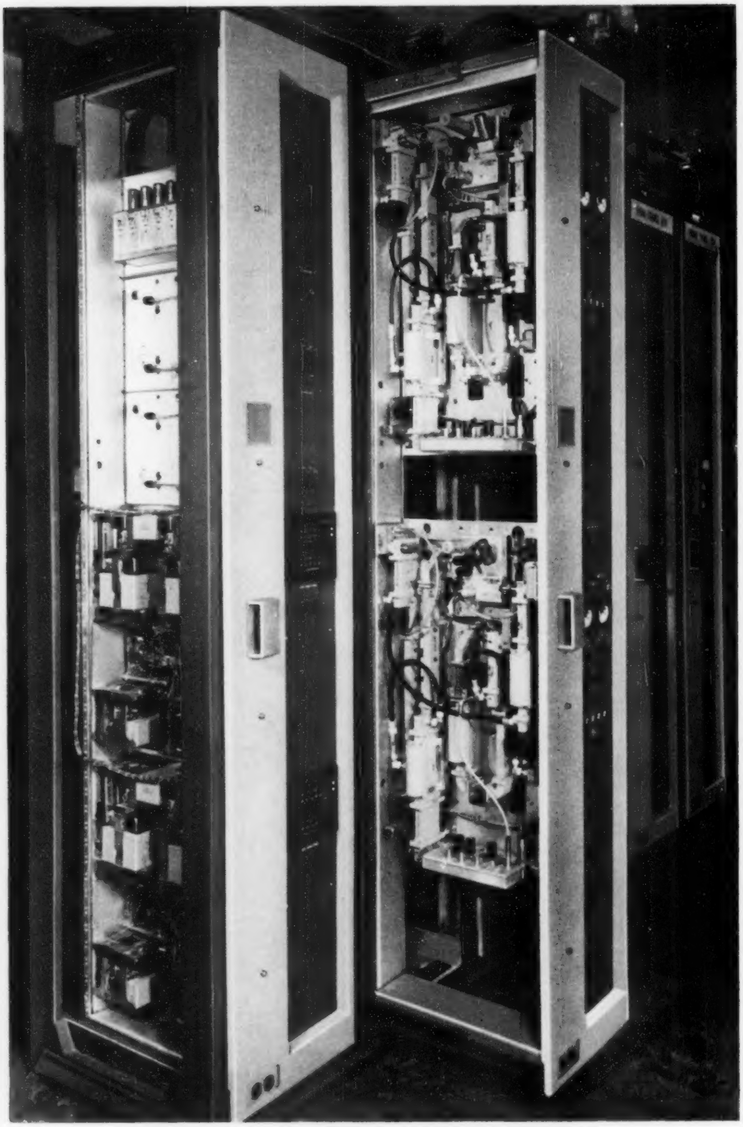

Fig. 8 — Bay arrangement for auxiliary radio channels.

pairs at an intermediate point, two bays of the type shown are required at such a station. When the station operates to one side only, two bays are required but are only half equipped. The channel separation networks for the auxiliary channels are not located on the panels but are placed in the horizontal waveguide runs in the space above the bays. Four waveguide connections serve two transmitters and two receivers in each bay. Since no more than four waveguide runs can be brought into one bay, the two microwave carrier supplies are brought in by minimum-length coaxial cables from transducers located immediately above the bay.

4.6 *Microwave Equipment Cooling*

In the radio transmitters and receivers and the microwave carrier supply, there are electron tubes requiring forced-air cooling to prevent overheating and subsequent shortening of tube life. The TWT amplifiers, in particular, must be provided with cooling air continuously. Loss of air on these tubes will result in severe damage to the tubes within a period of minutes and will disable the entire station. Accordingly, a

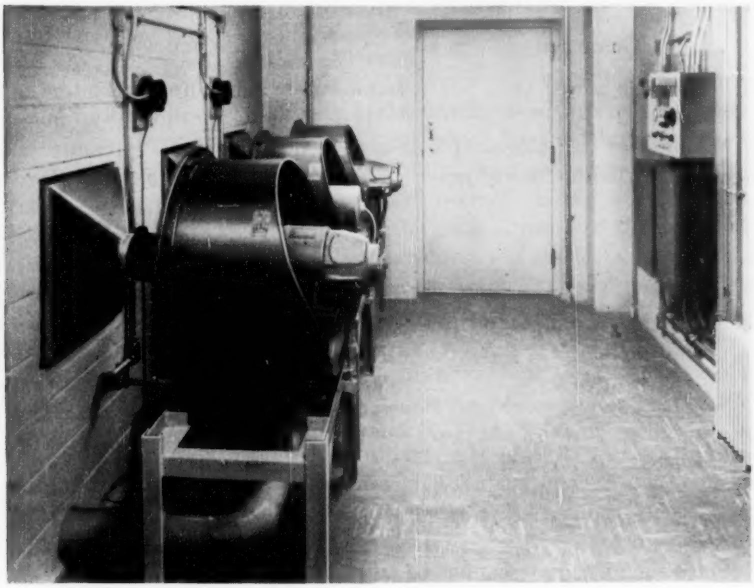

Fig. 9 — Typical blower system installation.

highly reliable central blower system supplies air to all of the transmitters and receivers and to the microwave carrier supply equipment for a full station of eight two-way channels. The capacity is adequate for maximum altitude and temperature conditions. As will be seen from Fig. 9, showing a typical blower system installation, three blowers are provided. These blowers are arranged to feed into a common duct assembly which, in turn, is connected to the duct provided in the base of the radio equipment bays. Since the spacing between the two radio bay line-ups has been standardized, a common blower duct assembly is used for all installations, thus realizing the economies inherent in the prefabrication of this assembly in the factory.

The use of three blowers and the power supply and control arrangements for these are the basis for the high reliability attained. In normal operation only one blower operates at a time, the back-flow through the idle blowers being blocked by automatic check valves in each blower output pipe. The primary blower is an ac powered machine which operates on the essential ac services supply for the TH station. The second and third blowers are dc machines which are driven, as ordered by the control circuits, from the 130V standby battery. The automatic control features perform the functions listed below.

Condition	Action
1. Loss of ac power	Immediately starts first dc blower and stops it on restoration of ac power.
2. Failure of ac blower for other than ac power failure	Senses reduction in air pressure and starts timing circuits associated with both dc blowers preparatory to starting. After approximately seven seconds starts first dc blower and halts timing for start of second dc blower.
3. Condition 2 plus failure of first dc blower	Completes timing sequence for second dc blower approximately 16 seconds after origin of condition 2 and starts second dc blower.

Alarms are reported for the conditions described above or for blower system failure. Only condition 1 is automatically reversible. Condition 2 or 3 locks the dc blower assigned in service until personnel can effect corrective action.

Since the air supply cannot be turned down after a station has been placed in service, special provisions have been made to permit the addition of bays to a line. Such additions are always at the end of a line away

from the blower supply, and gaps in a line are not permitted. The duct through the base of the bays has a movable plug at its outer end. As each bay is added, the movable plug is unlatched and advanced by the increment of one bay. Since the sealing action of the movable plug is imperfect, a gasketed cap is placed over the duct on the last bay of the line to seal the opening.

Temperature control of the radio equipment is part of the effort which has been made in all parts of the TH system to improve transmission stability. To this end, the room ambient where the radio equipment is located is held within limits which will not allow the temperature of a waveguide filter or network to deviate more than $\pm 10°F$ from the normal for that filter. The room ambient established is not necessarily critical, but it is expected that this will be within the range of 70° to 80°F. To meet this requirement, air conditioning is provided, in addition to the heating equipment normally furnished. Usually, the radio equipment is located in a separate room in order to reduce the air conditioning load requirements for the station. At stations where there is a need for air conditioning in the other areas, the radio equipment may be located in the same area with other equipment provided the temperature limits are satisfied.

V. FM TERMINALS

The FM terminals are designed for mounting on a conventional framework. This framework is the duct type which has been a Bell System standard for some years. The FM terminal with its power supplies on the 11-foot six-inch high bay framework is shown in Fig. 10. Also illustrated is the arrangement used for the plug-in units of the terminal. This general design is used for all active circuits of the terminal with the exception of the klystron oscillators in the transmitter. Insertion of the plug-in units automatically connects power and transmission leads and picks up cooling air when required.

The power supplies for both the transmitter and receiver are located at the top of the bay. Immediately below the power supplies is the FM transmitter, comprising six plug-in units and the klystron oscillator section. Jacks for access are located at the lower right of the transmitter. Below the transmitter appears a monitor and alarm control panel which serves as a mounting for the units monitoring the baseband amplifiers of both the transmitter and receiver. Provision is also made on this panel for alarm keys and lamps and appearances of test trunks.

The bottom portion of the bay is used for the FM receiver and a blower for cooling the equipment on the bay. The receiver requires three plug-in

Fig. 10 — FM terminal bay; photo at right shows plug-in arrangement.

units, when fully equipped, and has provisions for access jacks similar to those on the transmitter.

The use of the blower unit for each bay is a marked difference from the tube-cooling system for the microwave equipment. Several factors entered into this choice, foremost being the flexibility desired in location of FM terminal bays and the difficulty of engineering and installing air-duct work for a common air supply. Also, the reliability needed in the air supply for the FM terminal is not so critical since loss of the air supply will not result in immediate damage. With the arrangement

used, a supply pipe extends to the top of the bay from the blower unit. From this pipe, connections are made with plastic hose to the back of the units requiring cooling air.

VI. PROTECTION SWITCHING EQUIPMENT

Like the FM terminals, the protection switching equipment is mounted on 11-foot six-inch duct-type bay framework. Four types of bays are recognized, two of these being switching bays (baseband or IF) and the other two control bays (dc or tone control). The bays are always used in pairs consisting of one switching and one control bay, the manner of pairing being determined by the type of protection switching section

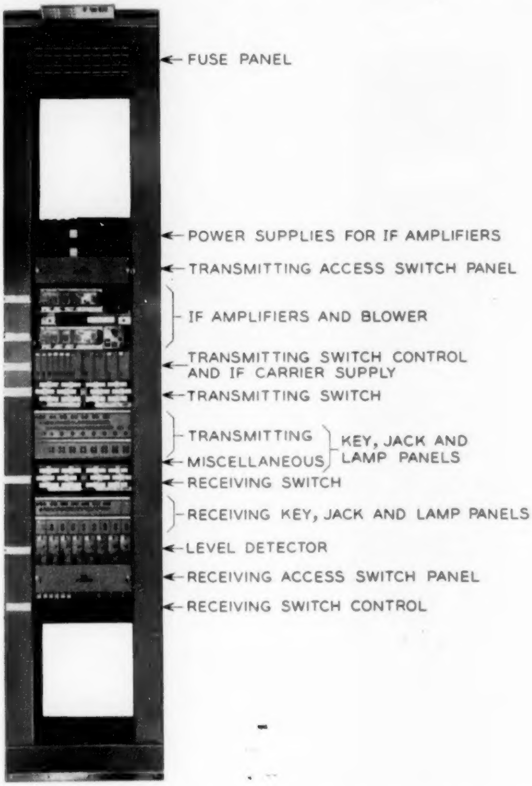

Fig. 11 — Intermediate-frequency protection switching bay.

to be served. Whatever the type of switching section, one pair of bays is required at each end and contains the equipment required to serve both directions of transmission within the section.

Four types of switching sections are possible, encompassing the allowable combinations of IF and baseband switching. Where FM terminals are involved, the choice of pairs of bays is dependent on whether the terminals are protected alone or in combination with radio links. This determines the method by which the ends of the switching sections communicate, one with the other. Where only FM terminals are protected, the switching section is baseband-to-IF (or IF-to-baseband), and both ends are located in the same building. Reports and orders between the ends are then provided by dc signals over station cabling. If radio links are included in the same switching section with FM terminals, the sec-

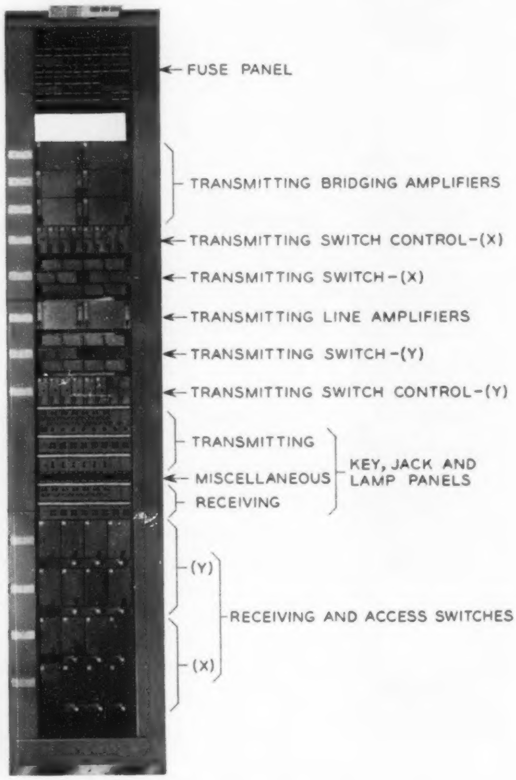

Fig. 12 — Baseband protection switching bay.

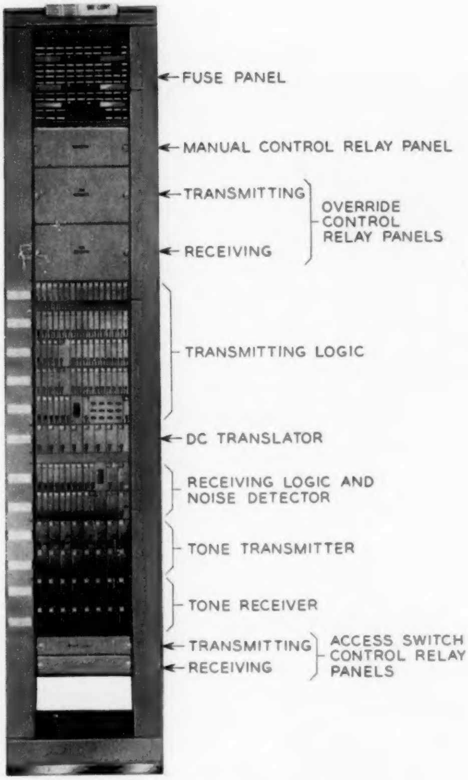

Fig. 13 — Protection switching control bay.

tion may be as above or it may include FM terminals at both ends with baseband-to-baseband switching. In either case, the communication between the ends is provided by a system of tones transmitted over the auxiliary radio channel. For protection of sections including radio links only, the same system of control tones is provided with IF switching equipment at both ends.

As implied above, the differences between the control bays lie in the equipment required for handling report and order signals and in the simplification of logic made possible with dc reporting. Thus, while the tone system requires a group of oscillators and detectors and a receive-end logic, the dc system employs dc reporter circuits, which can be wired directly to both ends of the section, and which reduce the receive-end logic to comparatively simple switch verification units.

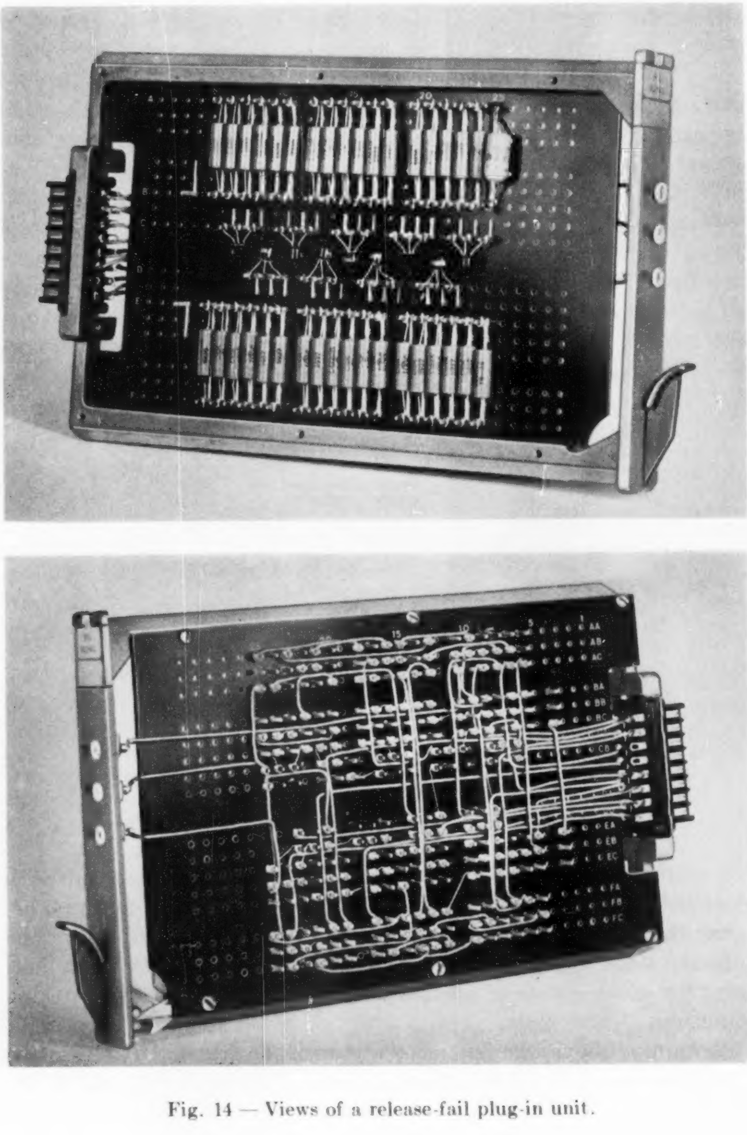

Fig. 14 — Views of a release-fail plug-in unit.

Three of the protection switching bays mentioned above are shown by Figs. 11, 12, and 13. The control bay for dc reporting is not shown since its general design is the same as that of its tone reporting counterpart. Examination of the bays reveals that there is a wide variation in the component units used in them. The more conventional are relay units and jack mountings. The IF switching bay also uses the same plug-in IF amplifier found in the FM terminal.

More than in any other equipment for the TH system, advantage has been taken of plug-in unit design for convenience in maintenance and additions. This is made possible by the extensive use of transistorized

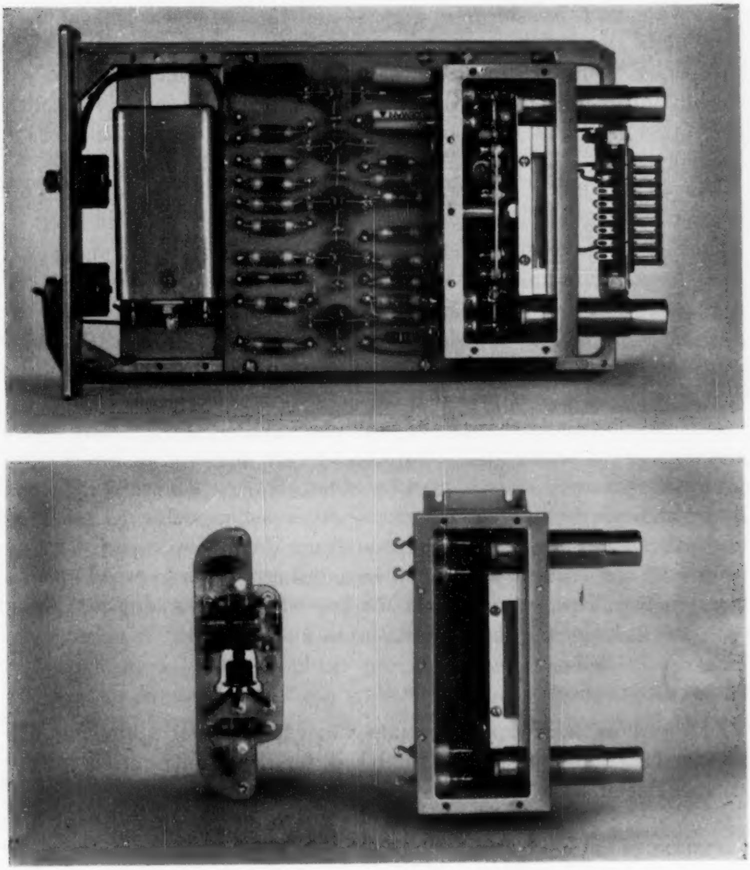

Fig. 15 — Views of an IF level detector unit.

circuits. For these, a basic design has been chosen which lends itself to a type of bookshelf mounting, wherein receptacles are affixed to the rear of the mounting compartment into which the units are plugged. The mountings are assembled such that the upper and lower members provide slider grooves at one-inch intervals horizontally and recesses into which latches on the units drop on insertion. The grooved parts are a single die-cast part which is used in all such applications.

The plug-in units are constructed on die-cast frames of which two basic sizes are used. The difference is in the face of the frames, which are available in nominal widths of one inch and two inches. Thus, one molding die with interchangeable slides for the face section can be used. Two of the units are shown in Figs. 14 and 15. The first is typical of the logic units and also several other units. The mounting board is a fiber part which is universally perforated to receive terminals designed specifically for the purpose. As many as 180 terminals can be inserted in the perforated grid. The components are mounted by their pigtail leads on one side of the board, and all wiring is placed on the opposite side.

The second unit is the IF level detector used in the IF switching bay (Fig. 15). In this design a two-inch face frame has been adapted for mounting circuitry involving a relay, as well as transistors and pigtail components. The unit also illustrates the introduction of coaxial connections to a unit while still maintaining the features of being completely plug-in. Other designs not shown make use of a pair of die-cast frames, with appropriate framework parts between, to build up units requiring greater volume than is provided by the two-inch face frame.

It should be noted that the bays illustrated are provided with all the plug-in units required for a complete protection group comprising two protection channels and six working channels. It is more often the case that the bays need be only partially equipped as delivered. In these cases, the unused positions of the mountings are filled with empty frames to protect the units installed and to avoid erroneous insertions during maintenance. The mountings on the bay are fully wired, and pairs of bays are fully interconnected for the ultimate growth.

VII. ACKNOWLEDGMENTS

The authors take this opportunity to express their appreciation to Messrs. G. H. Klemm and L. M. Klenk for their contributions to portions of this article.

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The TH Broadband Radio Transmitter and Receiver

By P. T. SPROUL and H. D. GRIFFITHS

(Manuscript received June 12, 1961)

A general description of the electrical and mechanical features of the TH radio system is given in the two preceding articles.[1,2] This paper describes the broadband radio receivers and transmitters in detail. Special attention is given to the new features: the receiver modulator with germanium diodes and its associated preamplifier with a noise figure of 10 db; the IF main amplifier in which nearly all adjustments have been eliminated; the amplifier-limiter with unusually low AM/PM conversion; the high-power transmitter modulator using a varactor diode to avoid conversion loss; and finally the 5-watt traveling-wave tube amplifier.

I. BROADBAND RECEIVER

1.1 *General*

Fig. 1 is a block diagram of the TH radio receiver. The incoming signal from the antenna waveguide system is selected by the channel separation network. After further filtering by the channel bandpass filter, the signal is applied to the receiver modulator. Here it is mixed with a beat oscillator (BO) frequency derived from the microwave carrier supply, to provide an intermediate frequency centered at 74.1 mc.

The IF signal is amplified first by a low noise preamplifier of the cascode type and then by the IF main amplifier. Under normal (no-fade) conditions the IF main amplifier provides about one-half of the gain of the radio repeater. Sufficient additional gain is available, as determined by the automatic gain control (AGC) circuit, to keep the receiver output level constant for repeater input signal levels ranging from 5 db above to 25 db below the normal. For short sections, of low path loss, a pad of suitable value (not shown in Fig. 1) is used between the preamplifier and main amplifier to keep the AGC circuit properly centered. Equalization follows the IF main amplifier. All receivers have a basic equalizer which compensates for gain and delay departures from

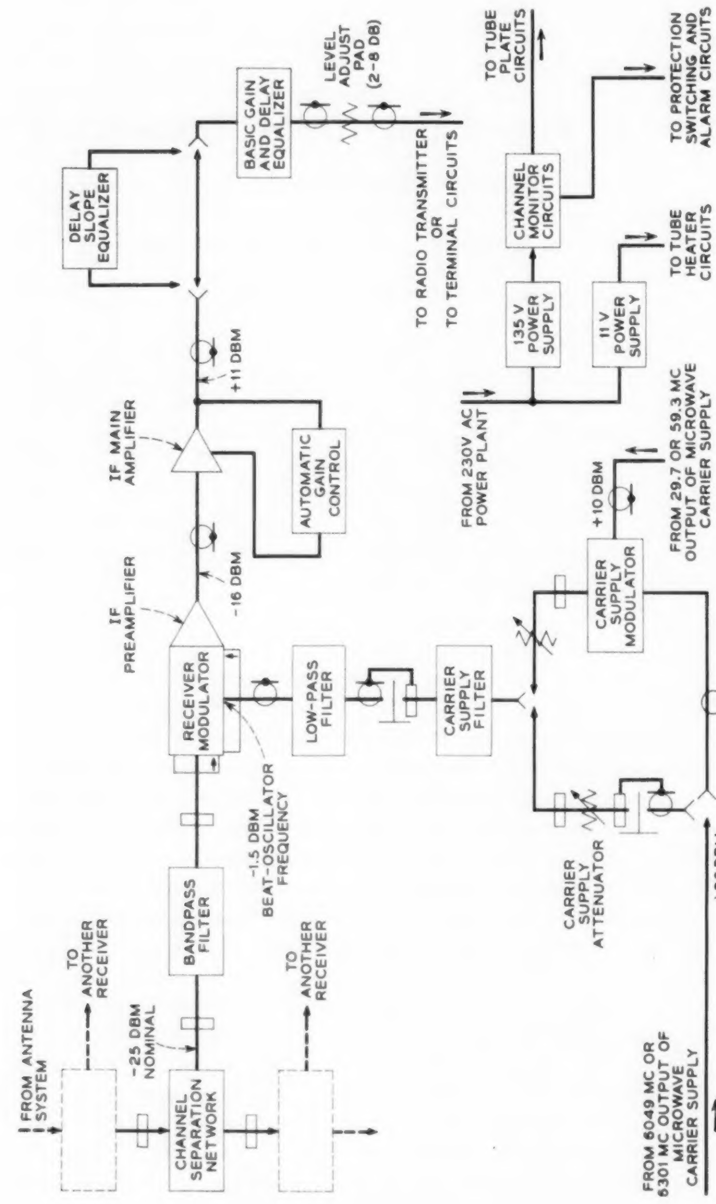


Fig. 1 — Block diagram of TH radio receiver.

ideal transmission for both the transmitter and receiver of the section. A delay slope equalizer is inserted between the IF main amplifier and the basic equalizer when required to correct for delay slope distortion which has accumulated in several repeaters. Not shown on Fig. 1 is a 93.8-mc rejection (trap) filter, connected between preamplifier and main amplifier and used only on channels 11 or 21 and 18 or 28; this filter blocks spurious transmission of the adjacent auxiliary radio channel through the broadband channel.

The nominal microwave signal power at the input to the channel bandpass filter is -25 dbm* and the output of the IF main amplifier is set for a constant level of approximately $+11$ dbm. The pad at the receiver output is chosen to give the proper power input at the connecting equipment.

The BO frequency of 6049 mc (or 6301 mc) is used directly for the receiver modulators of channels 12 and 17 (or 22 and 27). For the other channels, a carrier supply modulator is required as part of the radio receiver. This modulator shifts the 6049 mc (or 6301 mc) by either 29.7 mc or 59.3 mc. The carrier supply filter at the output of the modulator selects the desired BO frequency. However, this filter does not provide attenuation to the second harmonic output at 12 kmc from the modulator, which was found to be a source of interference. The low pass filter, following the carrier supply filter, attenuates these undesired frequencies. An attenuator is used to set the BO input power to the receiver modulator to the correct value for optimum noise figure of the crystals.

Indications of lack of input signal or malfunction of equipment are provided by the channel monitor circuit. This circuit informs the protection switching and alarm circuits when the input signal falls to approximately -52 dbm.

Fig. 2 is a photograph of the radio receiver. The multiposition switch and the connector immediately below it are used with an external meter to check the condition of the receiver's electron tubes and diodes while the receiver is in use.

1.2 *Receiver Modulator and IF Preamplifier*

A photograph of the receiver modulator and IF preamplifier is shown in Fig. 3. Since the signal level is at a minimum at this point in the receiver, the best possible signal-to-noise ratio is required, consistent with good transmission. The receiver modulator is built in a dual waveguide

* This value was chosen early in the TH development for design purposes. As discussed in Ref. 1, the currently accepted value is -27 dbm, based on an rms section loss of 64 db.

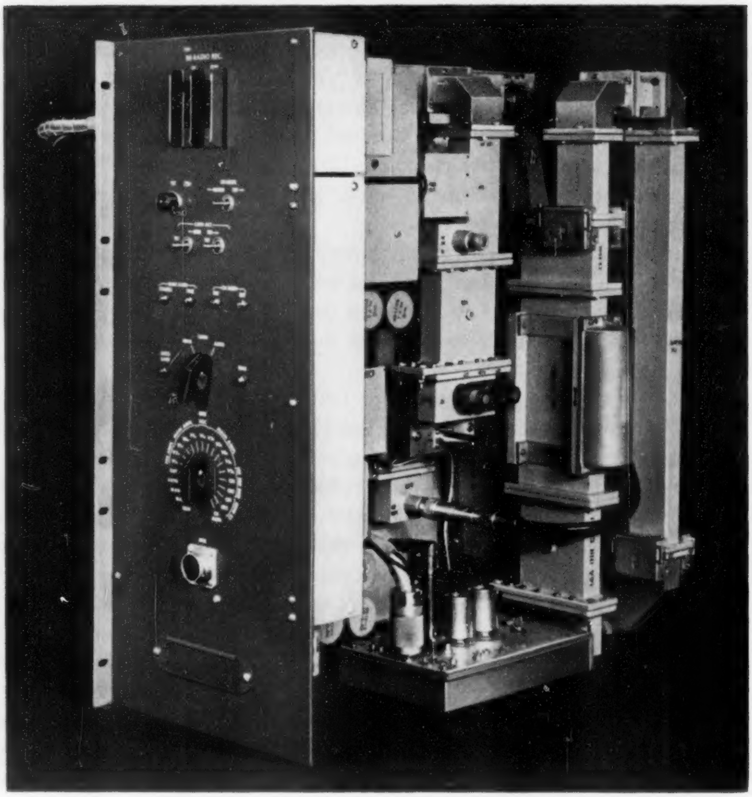


Fig. 2 — The broadband radio receiver.

configuration with four elements, the diode mount, the short slot hybrid junction, the dual isolator and the input transducer. It uses two germanium crystals in a balanced modulator circuit designed for optimum noise performance. The balanced circuit suppresses any amplitude modulation noise present on the BO input. The dual isolators absorb spurious frequencies generated in the modulator and provide a good input impedance. The IF preamplifier is a two-stage low noise amplifier of the cascode type using triode electron tubes. The noise figure for a typical receiver modulator and IF preamplifier with new crystals and tubes is about 10 db. The transmission is essentially flat over the band from 64 mc to 84 mc, dropping off by about 0.3 db at 58 mc and 90 mc.

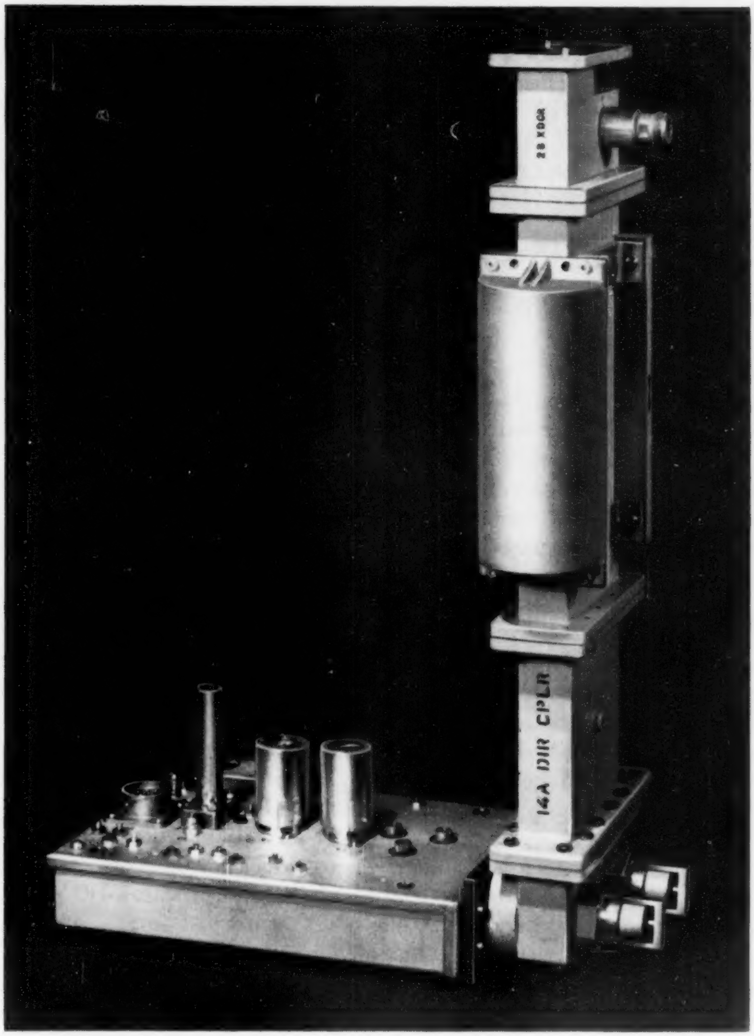


Fig. 3 — The receiver modulator (right) and IF preamplifier (left).

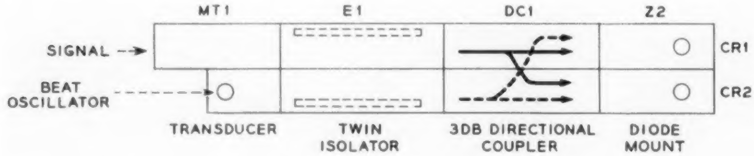


Fig. 4 — Simplified block schematic of the receiver modulator.

1.3 *Receiver Modulator*

A simplified block schematic of the receiver modulator is given in Fig. 4. The received signal is coupled through the transducer and one side of the twin isolator into the short slot junction which forms a 3-db directional coupler. The BO frequency is coupled to the directional coupler through the coaxial-to-waveguide portion of the input transducer and the other half of the twin isolator. The directional coupler divides the signal and BO powers equally and applies them to the crystals CR1 and CR2 of the diode mount.

A line drawing of the diode mount is shown in Fig. 5. The configuration is that of two coaxial-to-waveguide transducers placed side by side and sharing a common E plane wall, with the diodes in the transition region, i.e., partly in the waveguide and partly in the coaxial cavity. The depth that the coaxial sleeve penetrates into the waveguide controls the coupling between the coaxial line and the waveguide, thus matching the resistive portion of the diode impedance to that of the waveguide. The reactance is tuned out by the properly positioned fixed waveguide short circuit behind the junction and by adjustment of the length of the coaxial cavity, which also serves to compensate for differences between diodes. These latter adjustments, called RF TUNER 1 and RF TUNER 2 on Fig. 5, are the only mechanical adjustments.

The modulator uses germanium diodes, which are matched pairs of 1N263-type diodes. Optimum noise figure is obtained with a BO power of about -4.5 dbm and a fixed bias of about 0.3 volt from a 200-ohm source applied to each diode. The modulator has a conversion loss of approximately 6 db.

The IF circuit is designed to provide adequate decoupling to microwave frequencies and at the same time minimum capacitance. The decoupling is obtained through the use of an RF choke in the IF output lead. The choke consists of two one-quarter wavelength radial lines. One of these lines is resonant in each half of the 5925-mc to 6425-mc band. This gives a minimum insertion loss of 30 db over the band. The position of the choke along the IF output line is so chosen that a low

impedance is presented at the waveguide wall. Through the use of this construction, the total capacitance of the IF output terminals in parallel has been kept to 12 mmfd. Minimum capacitance is necessary because of the broad IF bandwidth to be transmitted.

The short-slot hybrid junction consists of two waveguides sharing a

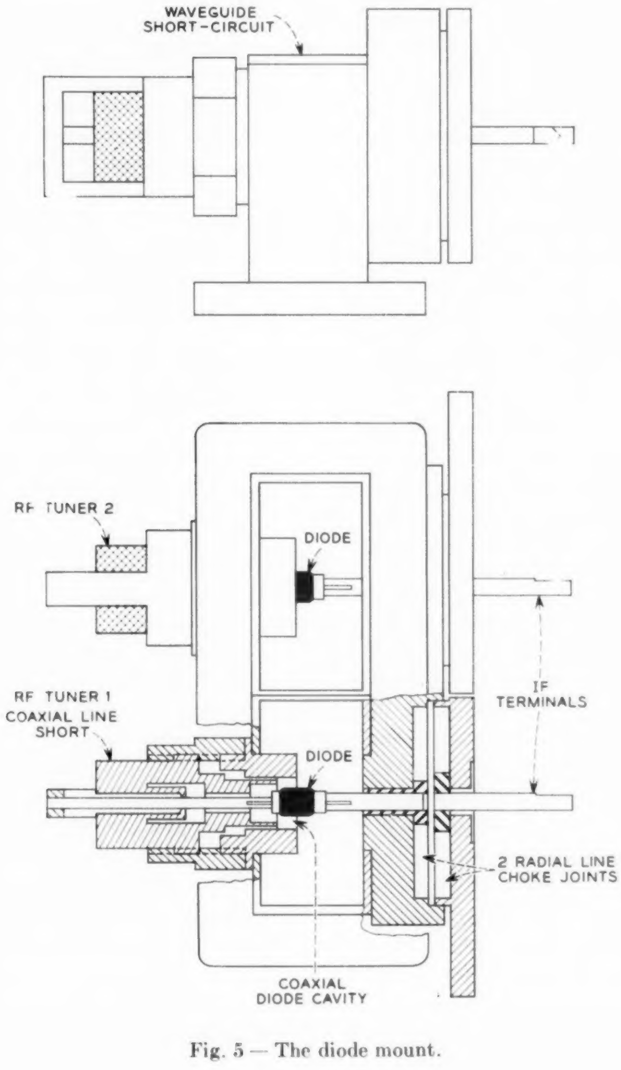


Fig. 5 — The diode mount.

common E plane wall in which there is a coupling slot to form a hybrid junction.[3] Power applied to one input arm divides equally between the two opposite output arms with no coupling to the third arm. Instead of the 0° or 180° phase difference between output arms as found in the hybrid T, the short slot coupler has a 90° phase difference. This causes the reflected signal power from the diode modulator to appear only in the BO arm rather than in the signal arm. Typical couplers have return losses greater than 24 db, a power split equal within 0.2 db, and a directivity of 23 db or better over the 500-mc band.

The twin isolator consists of two similar isolators mounted side by side with the waveguides having one common E plane wall. The isolators present a low loss to signals proceeding toward the modulator, and high loss to reflected energy.

One of the products generated by the modulator falls at twice the BO frequency minus the signal frequency. If this product is reflected back to the modulator in the correct phase, an improvement in the noise figure of up to one-half db can be obtained. However, the disadvantages of using this image frequency energy are more important. First, a different mechanical spacing of the channel bandpass filter would be required for each channel, and second, the variation in phase would cause the IF output impedance to vary over the band, which would adversely affect the transmission performance. An isolator, therefore, is provided in the BO arm to absorb the image as well as any signal energy reflected from the modulator.

1.4 *IF Preamplifier*

Fig. 6 is a schematic of the IF preamplifier circuit. It uses two 417A triodes connected in a cascode circuit between the IF terminals of the balanced modulator and the 75-ohm output coaxial line. The gain of the preamplifier is typically 15 db. As shown on Fig. 3, it is assembled in a shielded chassis about five inches wide, eight inches long and $1\frac{1}{2}$ inches deep.

The preamplifier is designed to provide a transmission characteristic which is stable over the life of the tubes, along with optimum noise performance. Factory and field adjustments have been minimized by the use of compensation circuits and by control of wiring in the transmission path. The compensation circuits are in the cathode circuit of the grounded-cathode first stage and in the grid circuit of the grounded-grid second stage of the cascode.

For a stable transmission characteristic, the input impedance of a tube, which is the termination of an interstage network, must be inde-

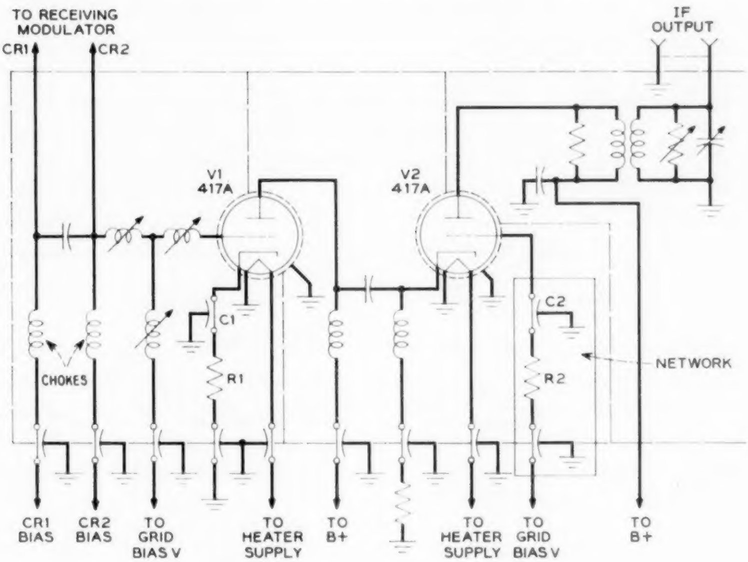


Fig. 6 — Simplified schematic of the IF preamplifier.

pendent of the voltage gain of the stage. For a triode with resistive load, the input capacitance C_{in} is given by

$$C_{in} = C_{gk} + (1 + A)C_{gp}$$

where C_{gk} is the hot grid-to-cathode capacitance, and A is the voltage amplification of the stage; C_{gp} is the grid-to-plate capacitance. Thus the input capacitance is a function of the voltage gain of the stage. This change in capacitance is commonly called the Miller effect. In many circuits this effect is cancelled by neutralizing techniques, one of which is to parallel the grid-to-plate capacitance with a suitable coil to form a parallel resonant circuit. Neutralization of this type could not be used because of the wide frequency band of 58 mc to 90 mc. However, resistance in the cathode circuit produces negative input capacitance, which also changes with the gain of the stage. By proper choice of resistor value (R1 of Fig. 6) the input capacitance is reduced and made, for practical purposes, independent of tube gain.

The lower curve of Fig. 7 gives the input resistance of a 417A in a grounded-cathode stage in which the connection to ground of the cathode pin of the tube socket is made as short as possible. This input resistive loading is due to cathode lead inductance and varies inversely with both

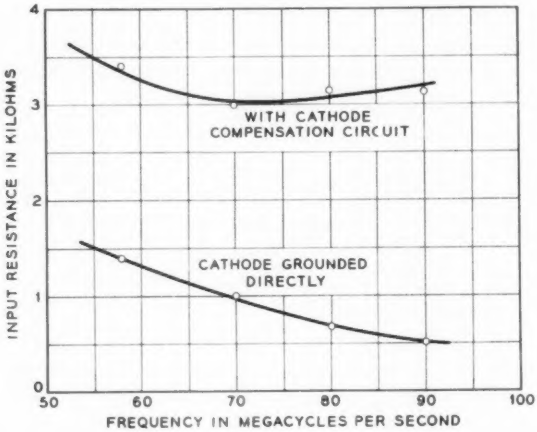


Fig. 7 — Input resistance of a W.E. 417A triode: cathode grounded directly and with cathode compensation circuit.

the square of the frequency and the operating gain. These variations in the input resistance make it a far from ideal termination for an interstage network. The inductance of the series resistor added for capacitance compensation makes the input resistance even lower. However, a capacitor in parallel with this resistor (C1 of Fig. 6) forms a broad series resonant circuit with the cathode lead inductance, and reduces the net inductance. The resultant input resistance, with the compensation circuit added, is plotted in the upper curve of Fig. 7.

Stability of the input impedance to the grounded-grid stage is obtained in a similar way by placing a small resistor and capacitor (R2 and C2 of Fig. 6) in parallel in the grid lead.

The coupling network between the receiver modulator and the grounded-cathode stage is an adjustable three-coil network in a T configuration. It is mismatched to obtain optimum noise performance, and is designed to operate with the IF impedance of the modulator as one termination and with the input impedance to the tube as the other. The only important resistive loading is applied by the real component of the IF impedance of the crystals.

The interstage network between the tubes appears as an extremely simple one on paper. However, stray impedances play an important role. The lead inductance of the connection between plate and cathode, the series blocking capacitor, the output capacitance of V1, and the input capacitance of V2 combine to form a filter with adequate band-

pass. This network is also misterminated, to obtain an optimum noise figure from the grounded-grid stage. Plate voltage for V1 and cathode voltage for V2 are applied through high impedance chokes. No adjustments are required since the lead inductance is carefully controlled during manufacture.

The output circuit uses a double-tuned transformer of the type discussed more fully in the next section of this paper. One capacitor adjustment is required to compensate for circuit variations. The output impedance is 75 ohms, to better than 20-db return loss.

1.5 *IF Main Amplifier*

The IF main amplifier operates at a gain of 27 db for the nominal receiver input signal level of -25 dbm, and has a maximum gain with new tubes of about 60 db. The gain-frequency characteristic for a typical IF amplifier at nominal gain setting is essentially flat over the 64-mc to 84-mc band, dropping at 58 mc and 90 mc by about 0.6 db. Changes in the transmission characteristic with change in operating gain of 25 db are typically within 0.5 db.

A photograph of the amplifier is shown in Fig. 8. The amplifier is assembled in a shielded chassis approximately $4\frac{1}{2}$ inches wide, $1\frac{1}{2}$ inches deep and 20 inches long. The electron tubes are covered by a common electrostatic shield. Cooling air from the bay is supplied to the tubes by means of a conduit mounted on the shield. Circuit elements are

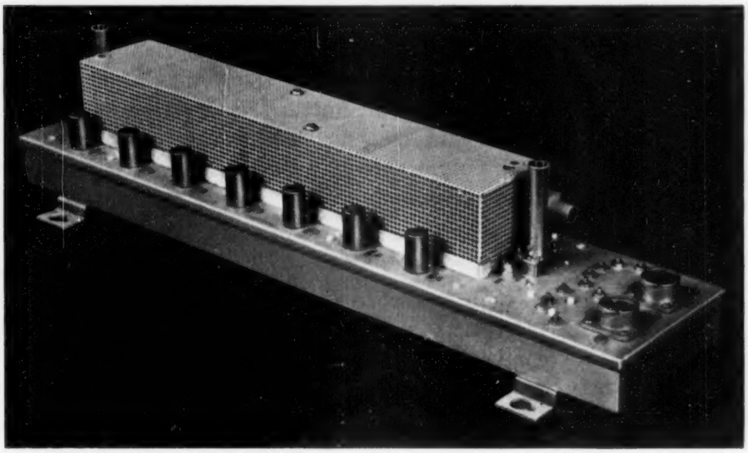


Fig. 8 — The IF main amplifier.

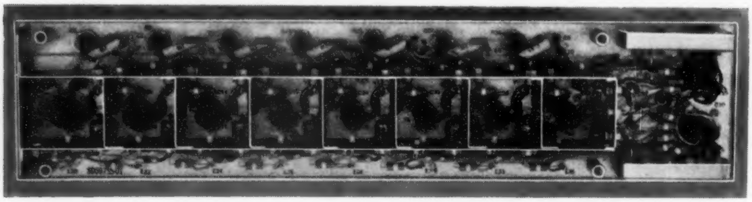


Fig. 9 — IF main amplifier, rear view with cover removed.

mounted on the underside of the chassis with each stage individually shielded, as shown by Fig. 9. Multi-pin connectors, which carry the necessary plate and grid supply voltages, also connect the individual cathode voltages to the control panel of the receiver for test purposes.

A simplified schematic of the amplifier is shown in Fig. 10. It uses a new high-performance tetrode, the Western Electric 448A, specially designed for this application.[4] Seven stages of amplification are used, the first, sixth, and seventh being fixed gain stages, the remaining four being variable gain stages. Fixed gain is used on the first stage to obtain the most stable input impedance, on the six and seventh to maintain the required power output over the range of operating gain. Connected across the output of the amplifier is a monitoring circuit which provides a dc voltage, dependent upon the signal level, to the automatic gain control (AGC) circuit. The latter controls the grid bias of the variable gain stages to maintain the output power constant as the input to the receiver changes due to fading, etc. By careful control of the tube parameters and of wiring during manufacture, by the use of compensation circuits, and by the use of an interstage transformer of a new type with a high coefficient of coupling, an amplifier has been achieved which requires no adjustment of gain-frequency characteristic when tubes are replaced. The input and output circuits each have two variable capacitors to permit optimum adjustment of impedance.

1.6 *Electron Tube Circuit*

The 448A tetrode has a transconductance of approximately 31,000 micromhos. It has very close spacings between the control grid and cathode and between the control grid and screen grid, and relatively large spacing between the screen grid and plate. This wide spacing makes possible a low in-circuit output capacitance, which averages 5 mmfd. The close control grid-to-cathode spacing makes negligible the effect of transit time on input impedance for frequencies up to at least 100 mc.

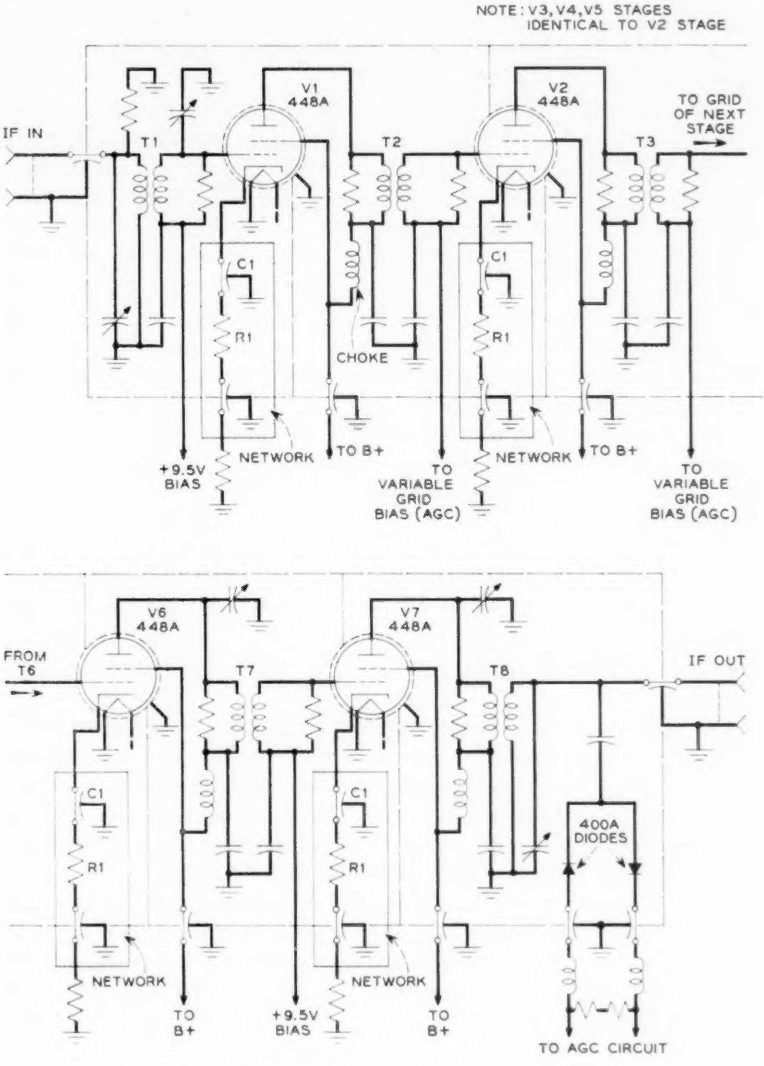


Fig. 10 — Simplified schematic of the IF main amplifier.

The input capacitance as measured in the circuit averages 23 mmfd. This tube therefore has a high gain-bandwidth factor, not all of which is realized in the amplifier since negative feedback is introduced to improve transmission stability.

The termination of an input or interstage network is the input impedance of the tube. For high transconductance tubes operating in the 50-mc to 100-mc region, this impedance is far from ideal. Cathode lead inductance gives rise to a positive input resistive loading, while screen lead inductance results in a negative input resistive loading. Both vary inversely with the square of the frequency and the operating gain. Furthermore, the input capacitance is a function of the total space current flowing past the control grid. Since the gain of several of the stages must be changed to provide AGC, the effect of these non-constant impedances must be minimized so that transmission through the stages will not be a function of the operating gain.

The required transmission stability is obtained by the use of compensation circuits in the cathode circuits of the tubes as described above for the preamplifier. The input capacitance of the 448A, as determined by measurement, is very nearly a linear function of transconductance over the operating current range for the gain-controlled stages. Resistance in the cathode circuit, while decreasing the stage gain, introduces a negative capacitance in the grid circuit, which also varies linearly with transconductance. With a suitable resistor in the cathode circuit, the input capacitance is stabilized at 23 mmfd over the operating gain range.

The more important contributor to the input resistive loading is the cathode lead inductance. To minimize this inductance, three cathode leads connected to widely separated base pins are provided in the tube. These are connected externally by a low inductance path consisting of a brass plate which joins together the appropriate tube socket pins. This results in an in-circuit cathode lead inductance of less than 0.01 microhenry. However, even this low inductance gives rise to an input resistance of a few hundred ohms at 90 mc at maximum gain. Fig. 11 shows measured values of the input resistance with the shorting plate directly grounded, as a function of frequency and cathode current. This input loading is comparable to the terminating resistance of the interstage on the grid side (approximately 110 ohms).

The resistor which provides compensation for input capacitance changes adds inductance to the cathode lead. However, by proper choice of a capacitor in parallel with the resistor, a compensation circuit was designed which reduces the effective cathode-to-ground inductance to about 0.005 microhenry over the 50-mc to 90-mc band.

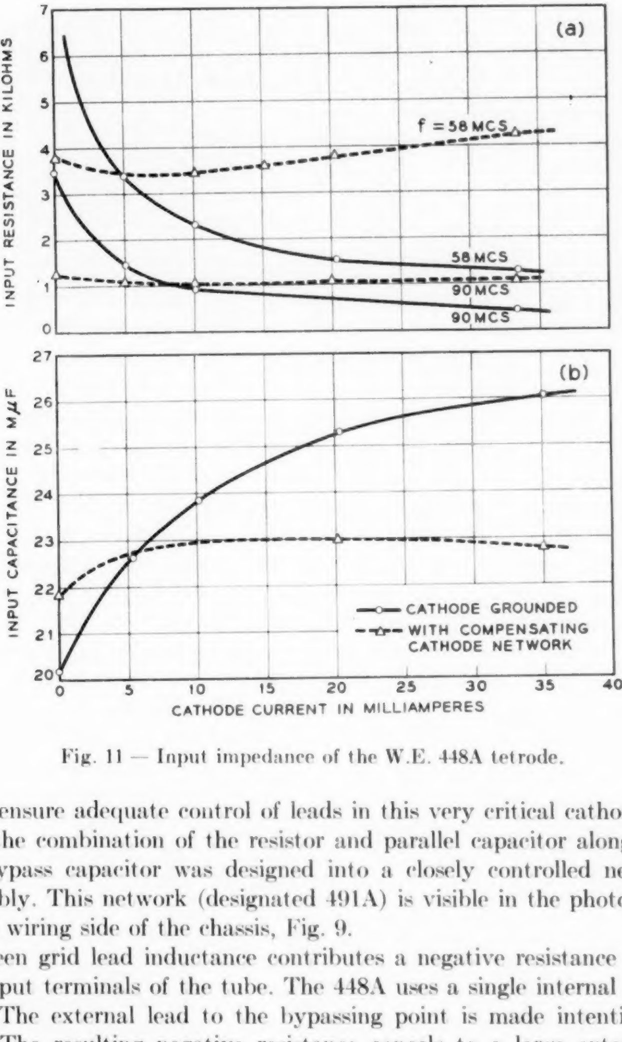


Fig. 11 — Input impedance of the W.E. 448A tetrode.

To ensure adequate control of leads in this very critical cathode circuit, the combination of the resistor and parallel capacitor along with the bypass capacitor was designed into a closely controlled network assembly. This network (designated 491A) is visible in the photograph of the wiring side of the chassis, Fig. 9.

Screen grid lead inductance contributes a negative resistance across the input terminals of the tube. The 448A uses a single internal screen lead. The external lead to the bypassing point is made intentionally long. The resulting negative resistance cancels to a large extent the loading due to the remaining cathode lead inductance. Fig. 11 shows the components of input impedance with the final compensation. The net effect is that of 23 mmfd shunted by approximately 1000 ohms. For all practical purposes, the resistance is independent of the value of cathode current, and for all frequencies is at least ten times that of the interstage terminating resistance.

All stages of the amplifier use dc feedback provided by a combination of self bias supplied by the cathode resistors and positive grid voltage obtained from a stabilized source. This minimizes gain variations from tube to tube and due to tube aging. The variable gain stages have less dc feedback than the fixed gain stages because of limitations imposed by the AGC circuit.

1.7 *Coupling Networks*

Three different networks are used in the amplifier: a terminated input network which couples from the 75-ohm input coaxial to the grid of the first tube, an interstage network, and a terminated output network which couples the plate of the last tube to the 75-ohm output coaxial line. Studies of various types of networks indicated that to meet the exacting requirements of minimum adjustment and permissible transmission deviations of a few hundredths of a db per stage, terminated networks would have to be used throughout. Of those studied, a design described by Rideout[5] which gives a slightly over-coupled characteristic and has the advantage of simplicity in design, is used. For the bandwidth required and the ratio of the terminating capacitances (5 mmfd for the tube output, 23 mmfd for the tube input) the network cannot be realized with a simple T or π configuration. However, it is realizable with either a tapped coil or a two-winding transformer. The disadvantage of the tapped coil is that a series blocking capacitor is required. For this reason the two-winding type was chosen.

The transformer is encased in an epoxy resin cylindrical block approximately $\frac{5}{8}$ inch in diameter by $1\frac{1}{8}$ inches high. The external view and stages in assembly showing the internal construction of a typical interstage transformer are shown in Fig. 12. Connections to the windings are made through four radial wires spaced at 90° intervals. The two windings are in separate grooves of a double thread cut in a plastic tube. The small mechanical separation between windings achieved in this way makes possible the required high coefficient of coupling without the use of a ferrite core. Because of the different impedance ratios required, the input, output and interstage transformers, although of the same general design, differ in a number of ways, such as the length-to-width ratio of the plastic tube, the number of turns and pitch of the windings, and the size of wire used. Externally, however, they are all similar in appearance. To achieve adequate uniformity in electrical characteristics, the individual inductances are held to less than ± 3 per cent of the nominal design value. Reproducibility of the transformers therefore depends upon very close control of all dimensions. The basic

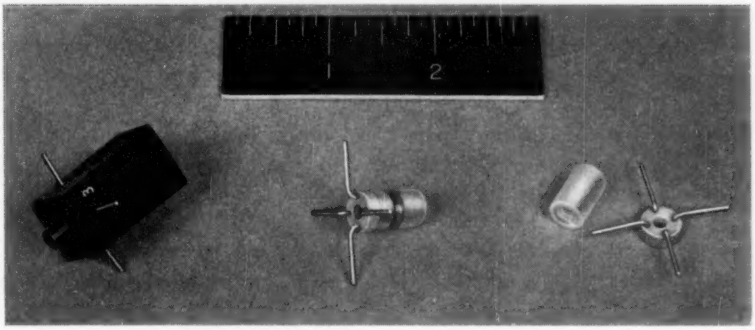


Fig. 12 — Typical interstage transformer.

transformer design has been so successful that it is used not only in the IF amplifier but throughout TH in broadband IF circuits.

To meet the objective of a minimum of factory and maintenance adjustments, a high degree of control over each of the components is required. Equally important is that their relationship to one another when installed in the amplifier remain the same from unit to unit. For this reason, the transformers are located in position by means of a keyed hole, and the interstage loading resistors are located uniformly. All components not directly in the IF transmission path are located outside the shielded area.

To minimize reflections (echoes) a good impedance match to the 75-ohm coaxial line is required at input and output. Typically the amplifiers have a minimum of 27 db return loss over the 64-mc to 84-mc band. To provide this performance consistently, two pairs of variable capacitors are added to correct for circuit variations. One additional variable capacitor is used in the sixth interstage. This capacitor is adjusted at the factory for optimum gain-frequency characteristic.

1.8 *Automatic Gain Control and Channel Monitor Circuits*

The AGC circuit is straightforward. A voltage doubler rectifier circuit at the output of the IF main amplifier monitors the signal power. Its output voltage, together with an adjustable reference voltage, is applied to the grid of a stabilized dc amplifier.[6] The reference voltage effectively sets the IF power at the main amplifier output. The amplified dc signal is applied to a cathode follower which translates it to the correct voltage level for control of the grid bias of the tubes in the IF amplifier. Thus an increase in IF power at the amplifier output results

in an increase in negative grid bias to the gain-controlled stages. A series resistance and shunt capacitance in the bias lead set the time constant and hence the response of the circuit to sudden IF signal changes. Diodes in the bias load limit the voltage rise when the IF main amplifier input is removed. The use of grid bias gain control is feasible as a result of the transmission stability obtained by the use of the compensating circuits in the cathode circuits of the tubes.

If the input signal is suddenly reduced by 27 db, approximately 30 milliseconds is required for the amplifier output to return to normal. This response time is adequate to follow the most rapid signal fading expected.

The channel monitor circuit initiates a local alarm and originates an order to the protection switching system when a deep fade of the incoming signal occurs. A fade results in an increase in the current drawn by the IF amplifier tubes. The increase in the voltage drop that this causes across a resistor in series with the plate supply is used to trigger a bistable transistor circuit. This in turn controls a relay to operate the alarm and call for a protective switch.

II. BROADBAND TRANSMITTER

2.1 *General*

The TH radio transmitter is shown in block schematic form in Fig. 13. The IF input signal from the radio receiver or terminal equipment is first applied to the amplifier-limiter. The limiter effectively removes any amplitude modulation that might be present in the input signal and which could be converted into phase modulation by the traveling-wave tube (TWT) amplifier. Such modulation would appear in the system as unequalizable distortion. Associated with the amplifier-limiter is the carrier resupply circuit, which provides a local carrier source if a deep fade or an equipment failure occurs.

The IF signal from the limiter output passes with negligible loss through the carrier resupply to the transmitter modulator. There, after passing through a buffer amplifier, the signal is mixed with the BO frequency in a low-loss balanced modulator. After selection of the appropriate sideband by the channel bandpass filter, the signal is applied to the transmitter amplifier. This TWT amplifier provides about 32-db gain and an output power of +37 dbm (5 watts). A filter following the amplifier attenuates its second harmonic output an additional 25 db. Next is an isolator to provide an adequate termination for the channel separation network. After passing through the power output monitor,

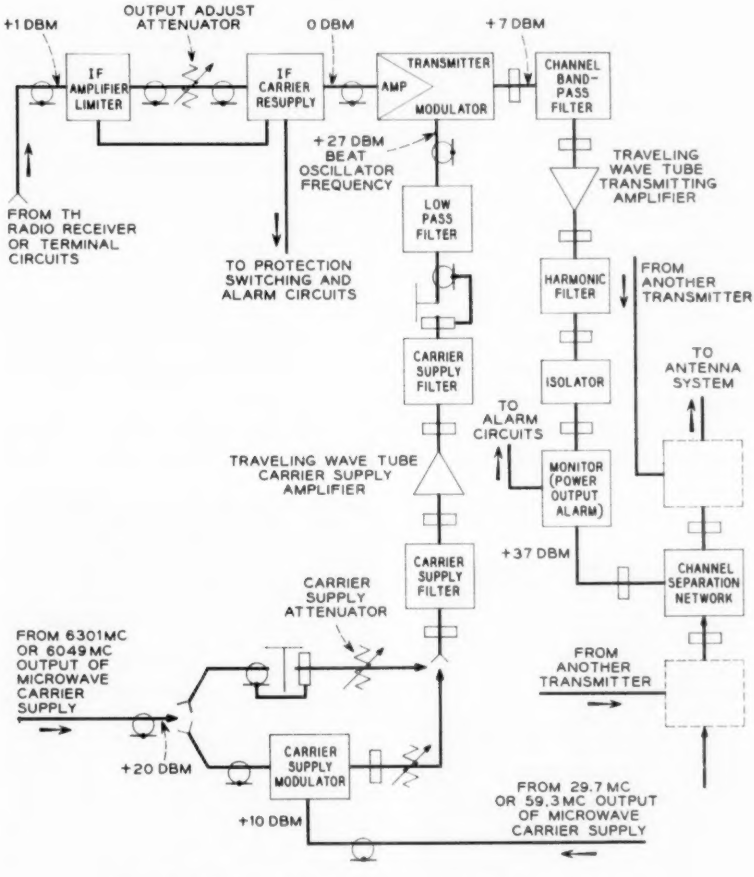


Fig. 13 — Block schematic of the TH radio transmitter.

the signal is fed to the channel separation network, which combines it with the outputs of other transmitters for transmission to the antenna system.

The power monitor circuit gives an indication to the repeater alarm circuits if the power drops more than 4.5 db. It consists of a 23-db directional coupler in the signal path feeding a crystal detector whose output current controls a meter-type relay. The meter also provides a visual indication of the output power.

The BO frequency for the transmitter modulator is obtained from the microwave carrier supply, as already described for the receiver.

However, here a relatively high power +27 dbm (0.5 watt) is required. As shown in Fig. 13, a second TWT amplifier, of the same type as used in the transmission path, is used to provide this power. Although this single-frequency, low-power application of the transmitting amplifier appears inefficient, it avoids introducing a second type of microwave amplifier into the repeater and permits the use of a common power supply for both amplifiers.

The carrier supply filter is divided into two sections, one ahead and one after the TWT. The combination provides the same amount of filtering as in the receiver. The section at the input of the TWT attenuates the unwanted sideband sufficiently to prevent cross-modulation in the TWT. The section at the output attenuates noise at signal frequency generated by the TWT amplifier. The low-pass filter attenuates second harmonic energy, as in the receiver.

Fig. 14 is a photograph of the radio transmitter. The main items visible on the front panel are the access doors to the traveling-wave tubes,

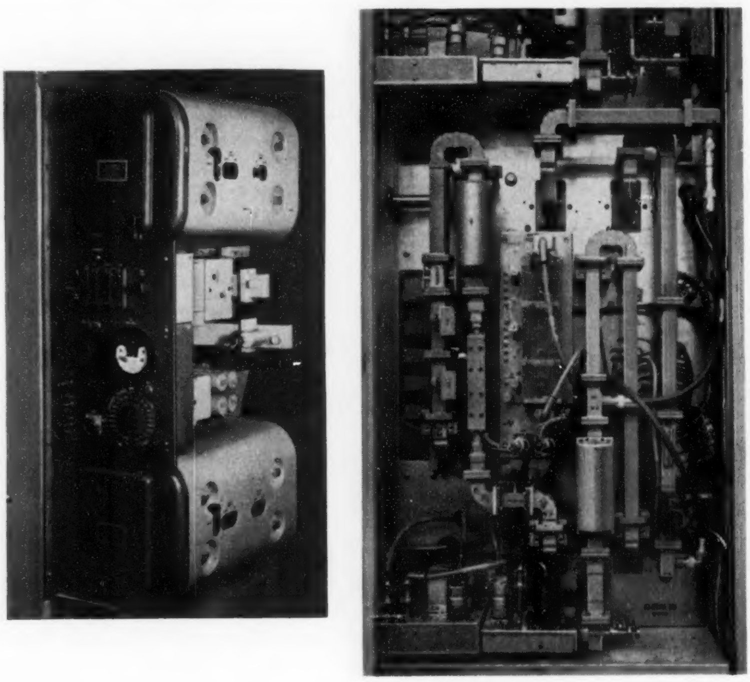


Fig. 14 — The broadband radio transmitter, and left side view.

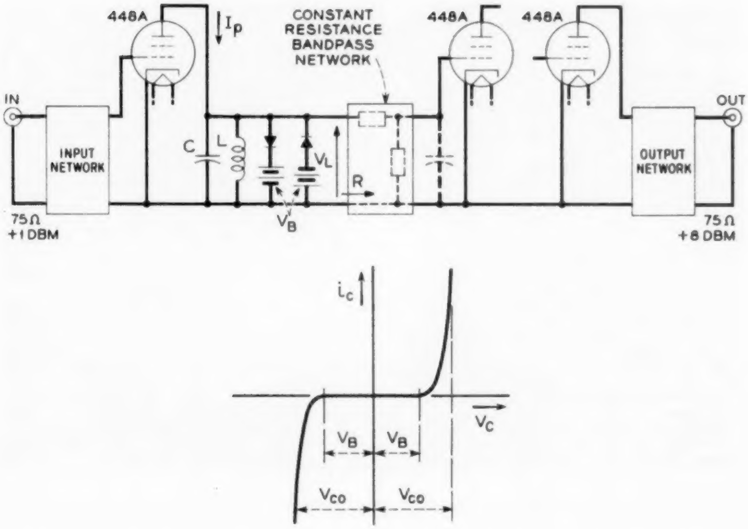


Fig. 15 — Simplified schematic of the IF amplifier-limiter, with typical voltage-current dc clipping characteristic.

the attenuator used to set the output level to +37 dbm, the output monitor meter, and the multi-position meter switch.

2.2 *Amplifier-Limiter*

A simplified schematic of the amplifier-limiter is shown in Fig. 15 and a photograph in Fig. 16. It uses three 448A high-transconductance tubes with compensation circuits identical to those in the IF main amplifier. To provide the required limiting, two 431A gold-bonded germanium diodes are used in each of the two identical interstages. The interstages are designed so as to provide maximum compression with a minimum of amplitude modulation to phase modulation (AM/PM) conversion. For this reason, the diodes are arranged in a symmetrical voltage clipping circuit in a parallel RLC circuit resonant at approximately mid-band. The diodes work into a constant resistance bandpass network. This network transforms the capacitance at the input to the following tube to a resistance of about 150 ohms in shunt with a small capacitance across the diodes. Harmonic frequencies generated by the diodes lie outside the bandpass of the network and are absorbed. The input and output networks are similar to those used in the IF main amplifier. Diode monitors (not shown) which produce a voltage proportional to

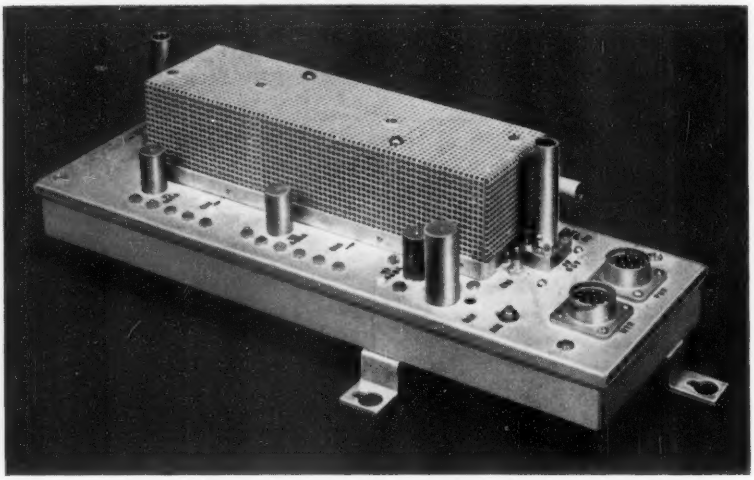


Fig. 16 — The IF amplifier-limiter.

the signal level are used at both the input and output. The voltage from the input monitor is used to operate the carrier resupply circuit, which is switched on when the signal power at this point drops 3 db. Both monitors are connected to the metering circuits to provide a convenient means of level checking.

The gain through the amplifier-limiter is normally about 7 db. When the incoming carrier is lost, the limiter acts as a linear amplifier with about 22 db of gain. Operation of the carrier resupply circuit provides a bias to the three tubes which reduces this gain to a loss of 2 db.

2.3 *Interstage Networks*

The design of the interstage networks is a compromise among maximum compression, minimum AM/PM conversion and a suitable gain-frequency characteristic.

The compression, C, is defined as

$$C = \frac{dV_i/V_i}{dV_0/V_0} \tag{1}$$

where V_i is the input voltage and V_0 is the output voltage. Compression is expressed in db by taking 20 log C. Fig. 15 shows a typical voltage-current characteristic of a symmetrical voltage clipping circuit. The factors affecting the compression of such a circuit are the voltage-

current characteristic of the diodes, the peak current supplied by the source, and the clipping level, which is dependent upon the bias supplied to the diodes.

The 431A gold-bonded germanium diode has a low forward resistance and good high-frequency characteristics. At 10 ma forward current, the ac resistance is between 4 and 5 ohms. A nearly ideal characteristic can be obtained with a low bias voltage of approximately 0.7 volt.

The dc clipping characteristic shown in Fig. 15 can be approximated by a power series containing odd powers only and for practical purposes, in the clipping region, can be reduced to one term, namely $i_c \sim V_c^n$, where n is an odd integer. The exponent n increases with diode bias, i.e. as the characteristic becomes more nearly ideal. The maximum compression, C, of the double-diode circuit, as calculated by nonlinear circuit analysis, is given by

$$C = n. \tag{2}$$

A value of $n = 15$ gives the best fit for the characteristic using 431A diodes. This means that 23.5 db of compression should be available from this stage. However, actual measurements show only 16 db. The main reason for this discrepancy lies in the over-simplified model for the diodes. Minority carrier storage in the diodes prevents current and voltage from following the dc diode characteristic at high speeds. Lifetime of the carrier in the 431A diode is relatively high, between one and two microseconds. Other types of diodes, such as the microwave diodes used in down converters, have negligible storage effects, but the amount of limiting attainable is small because of the rather high forward resistance of these diodes. High-speed computer diodes also exhibit very small storage effects (lifetimes of one to two millimicroseconds). Experiments with such diodes showed that higher compression could be obtained, but the capacitance of the diodes was rather high and AM/PM conversion, as explained later, would have been adversely affected.

In the Appendix, a simplified analysis of the circuit of Fig. 15 shows that the AM/PM conversion, P, is given by

$$P = 106 \frac{\Delta f c V_{co}}{I_p} \text{ degrees/db} \tag{3}$$

where

Δf = frequency deviation of the signal from center frequency,
V_{co} = voltage at which clipping occurs,
c = shunt capacitance across the diodes, and
I_p = peak current into the diodes.

This expression shows that the capacitance and the clipping voltage

should be small while the peak current delivered should be large, to keep the AM/PM conversion small. Exact tuning of the limiter circuit to the incoming carrier ($\Delta f = 0$) would seem to eliminate AM/PM conversion. However, a more exact solution, taking into account higher harmonics, indicates that there exists no frequency at which P is zero for all values of I_p. Measurements taken on the complete amplifier show this. (It should be noted that most of the AM/PM conversion is produced in the first stage, since the compression of the first stage reduces the AM input to the second stage.) The clipping voltage is kept small by the use of the low forward resistance 431A diode, and a high peak current is obtained by using a high transconductance tube, the 448A, as the driver. The shunt capacitance across the diodes is kept low by the network between the diodes and the following tube. On the tube side, the network is terminated by the input capacitance of the tube, namely 23 mmfd. To the diodes the network presents a resistance of 150 ohms, constant within 10 per cent, and a parallel capacitance of about 1 mmfd over a frequency range of 40 mc to 200 mc. Since the plate capacitance of the 448A is only about 2 mmfd, the use of a con-

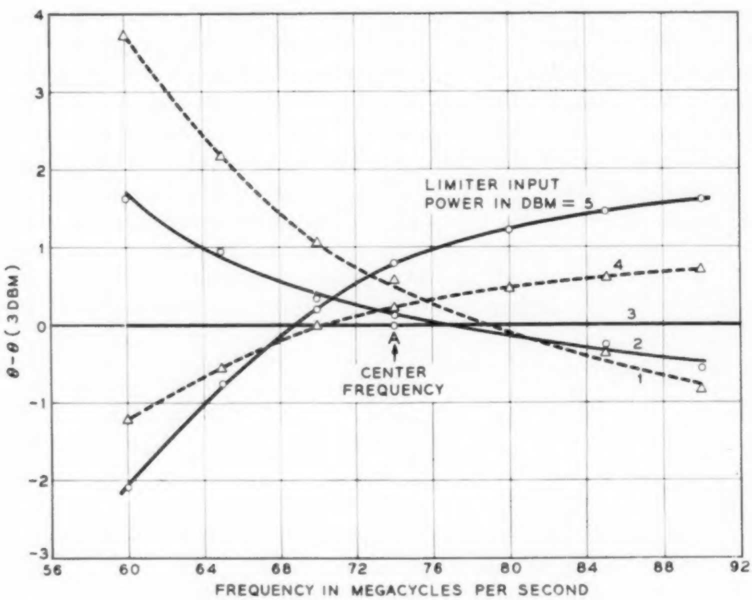


Fig. 17 — AM to PM conversion as a function of frequency with limiter input power as a parameter.

stant resistance network to reduce this capacitance is not justified. The capacitance of the two 431A diodes is only 0.9 mmfd when measured with small signals at 1 volt reverse bias. However, when measured in the circuit under heavy clipping, it is as high as 7 mmfd. This effect is again caused by minority carrier storage effects. The total capacitance across the circuit at high levels is approximately 10 mmfd. With $V_{co} = 1.1$ volts, $I_p = 15$ ma, substitution in (3) gives

$$P = 0.078 \, \Delta f \text{ deg/db for } f \text{ in mc.} \qquad (4)$$

For $\Delta f = \pm 15$ mc, the expected P would therefore be $0.078 \times 30 = 2.3$ deg/db. This can be compared with the measured results shown in Fig. 17. The curves show a total phase change from 60 mc to 90 mc of 1.9° at +4 dbm input power and 2.1° at +2 dbm input power, compared to a +3 dbm reference power. The very simple limiter model, therefore, gives results which agree well with measurements.

Fig. 17 shows the phase curves for the different input powers intersecting each other at certain frequencies. The intersection can be shifted in frequency by changing the parallel inductance L in the diode circuit. In this way it is possible to minimize AM/PM conversion for the carrier frequency of 74.1 mc at the nominal input power to the limiter. Fig. 18 shows the phase shift at center frequency as a function of input power. The AM/PM conversion is zero at +3-dbm input power but increases rather rapidly for higher and lower powers. It stays below 0.2 deg/db (or −30 db) over a range of more than 1.5 db. The input power to the limiter is controlled by the AGC of the receiver to less than ±0.5 db variation about the nominal power, so that the AM/PM objective is always met. (The measurements of Figs. 17 and 18 were based on a

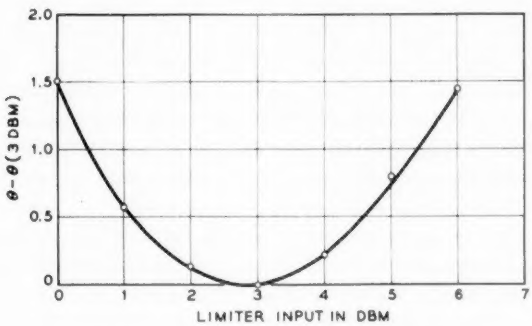


Fig. 18 — AM to PM conversion at 74 mc as a function of limiter input power.

nominal input power to the limiter of +3 dbm. This power was later lowered to the present value of +1 dbm. The inductance L was changed in value to optimize the circuit at the new power).

The limiter has to operate over a frequency range from 58 mc to 90 mc with only a few tenths of a decibel drop in transmission at the edges of the band. The basic diode network of Fig. 15 is inherently very broadband due to the limiting process. Its bandwidth is $B = f_0/Q$, and using (7), (8), and (9) we obtain:

$$B = \frac{I_p}{8cV_{co}} \tag{5}$$

which gives 131 mc if the values used for (9) above are inserted. The constant resistance networks have a bandpass characteristic which is flat to 0.01 db over the whole band. They also serve as harmonic filters, especially for the fairly strong third harmonic generated by the diodes. The over-all transmission characteristic of the limiter is flat to ±0.05 db from 64 mc to 84 mc and drops not more than 0.2 db at 58 mc and 90 mc.

2.4 *Dynamic Characteristics*

The preceding discussion considers only static amplitude changes. In practice the amplitude changes may have rates of up to 10 mc or even higher. Because of the large bandwidth of the diode circuit, the dynamic situation is little different. An approximate nonlinear analysis of the dynamic case indicates that the equations for compression and AM/PM conversion have to be multiplied by

$$\left[1 + \left(\frac{2f_m}{B}\right)^2\right]^{-\frac{1}{2}}$$

where f_m is the modulating frequency and B is given by (5). For f_m = 10 mc, this factor has a value of 0.988 and can therefore be neglected.

Fig. 19 shows dynamic compression in db for the complete two-stage limiter as a function of the modulating frequency. The curve is very flat with frequency, although the drop is somewhat greater than indicated by the theory. This compression curve can be affected by the impedance of the diode bias supply, through detection and remodulation. Such effects are avoided by the use of a very low resistive impedance bias source. A temperature compensation circuit changes the bias for the diodes in the second limiting stage to counteract temperature effect on the forward voltage drop of the diodes. In this way the output power of the limiter is constant to 0.05 db over a temperature range from 10° to 40°C.

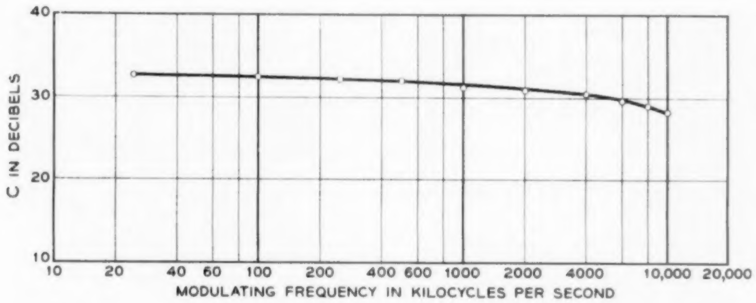


Fig. 19 — Dynamic compression for two-stage limiter as a function of modulating frequency.

The dynamic AM/PM conversion of the complete limiter was also measured. The accuracy of such measurements is affected by the difficulty of generating a 74-mc AM test signal free from PM. The results will include the AM/PM conversion generated by the unsymmetrical delay characteristic of the passive networks in the limiter circuits as well as AM/PM conversion due to limiter action. The measurements indicate the over-all dynamic AM/PM conversion is better than −30 db up to a modulating frequency of 10 mc.

2.5 *IF Carrier Resupply*

Without remedial measures, loss of the carrier allows the limiter and main IF amplifier to go to maximum gain. Within two repeater sections, the TWT amplifiers are saturated at full power with noise spread over a wide band. In the adjacent channels, this noise power is intolerably large. Protection against this is the function of the carrier resupply. When a carrier is lost, the gain of the limiter will reach maximum in a few millimicroseconds and the gain of the main IF amplifier in a few milliseconds. The carrier resupply replaces the lost carrier rapidly (within less than 0.1 millisec) and prevents subsequent repeaters from going to maximum gain. In addition, the limiter gain is reduced to attenuate the incoming noise and prevent it from modulating the new carrier.

The carrier resupply unit is a fully transistorized circuit containing a dc amplifier, a trigger circuit, a crystal-controlled oscillator, an amplifier, and circuits to reduce the limiter gain and operate a relay. The relay actuates the alarm and the automatic protection switching circuits. The ac equivalent circuit of the oscillator and amplifier is shown in Fig. 20; the other circuits, being conventional, are not shown. In the

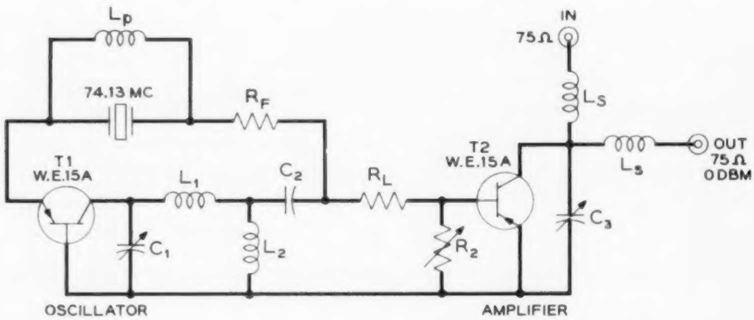


Fig. 20 — The ac equivalent of carrier resupply IF oscillator and amplifier.

oscillator and amplifier stages, high-frequency transistors having an alpha cutoff frequency of 750 mc are used. When a carrier has to be supplied, the transistors T1 and T2 are switched from cutoff to the operating point by the trigger circuit. The crystal oscillator starts oscillating in 60 microseconds. This rapid start time is obtained by resistor R_F in series with the crystal which lowers its Q. The crystal operates in the series-resonant mode at 74.1 mc, and the inductor L_p tunes out the parallel crystal capacitance.

The crystal is in the feedback loop of a grounded base transistor. The current gain necessary to produce oscillations and feed power into the load R_L is provided by the circuit $C1$, $L1$, $C2$, $L2$, which acts as an ideal transformer at the oscillating frequency of 74.1 mc. The second transistor T2 is connected as a grounded emitter amplifier. The current gain of this stage is approximately 15 db at 74 mc. The collector of T2 is bridged across a 75-ohm coaxial line by means of an extremely wide band low-pass filter consisting of L_S, $C3$, L_S. $R2$ adjusts the carrier power delivered to the 75-ohm output.

Measurements show the frequency of the oscillator to remain within ± 10 kc over a wide temperature range. This is well within the required limit of ± 50 kc. The starting time of the oscillator is dependent on temperature, but the variation is only a few microseconds over the range expected in the temperature-controlled radio repeater rooms.

2.6 *Transmitter and Carrier Supply Modulators*

The conversion gain available from the use of variable capacitance diodes as up-converters is important in the performance of the high-power microwave modulators used in the TH system. The transmitter modulator is used in the radio transmitter to convert the frequency

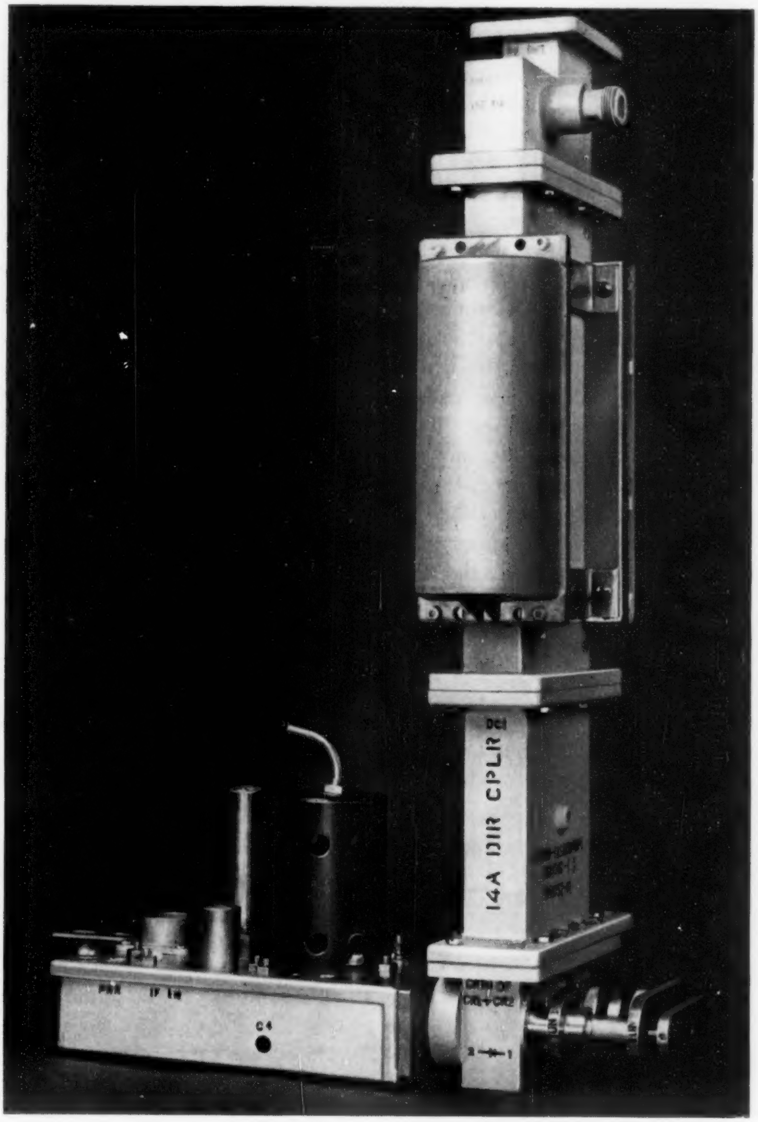


Fig. 21 — Complete transmitter modulator.

modulated IF signal of 74.1 mc to the required microwave frequency. Transmission performance is important in this modulator, and the diodes are biased to give approximately unity gain in order to obtain wide bandwidth. A photograph of a complete transmitter modulator is shown in Fig. 21. The carrier supply modulators are used in the microwave carrier supply to shift a microwave carrier by 252 mc, and in the radio transmitter and receiver to shift a microwave carrier by 29.7 mc or 59.3 mc. These are single-frequency devices, and conversion efficiency has been optimized.

The two types of modulators use a similar double-waveguide type of diode mount, and are shown schematically in Fig. 22. The diode mount provides a means of matching the 427A diode to WR159 waveguide, and is shown in Fig. 23. It is the same in principle as, but differs in detail from, the receiver modulator diode mount of Fig. 5. To match the resistive component of the diode impedance to the characteristic impedance of the waveguide, the coaxial line in which the diode is mounted is placed an appropriate distance off center in the waveguide. To compensate for differences between diodes, the coupling between the diode and the waveguide is adjusted by controlling the distance that the outer conductor of the coaxial line extends into the waveguide (RF TUNER 1 of Fig. 23). Reactance at the junction is partially cancelled by a waveguide short circuit fixed in position behind the coaxial extension. The reactance that remains is tuned out by means of a short circuit placed in the coaxial line. In the case of the transmitter modulator, this short circuit is made adjustable to compensate for differences between diodes (RF TUNER 2 of Fig. 23) and is set for optimum transmission over the band. In the carrier supply modulators, the position of the short circuit is fixed.

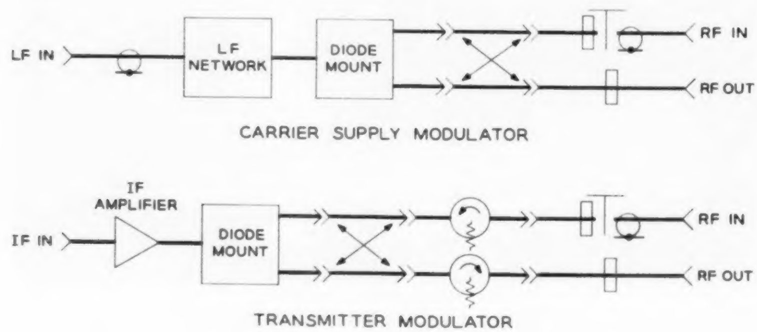


Fig. 22 — Schematics of the two types of modulators.

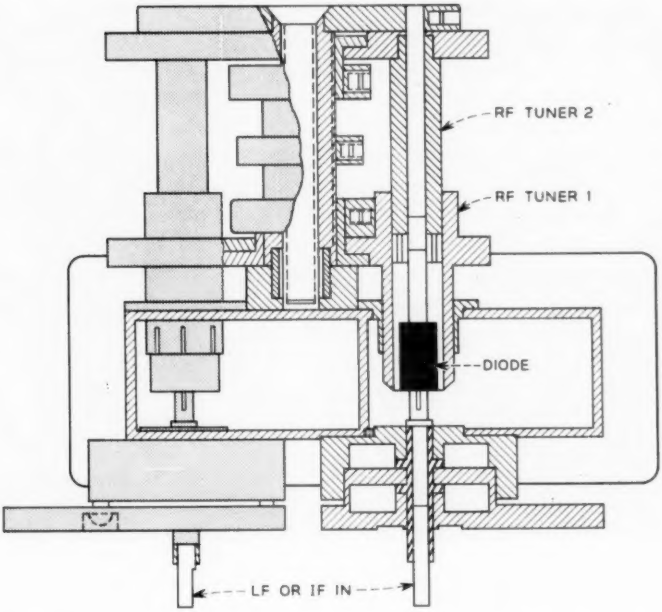


Fig. 23 — Double diode mount for transmitter and carrier supply modulator.

The use of radial line RF chokes in the input leads, a short-slot hybrid junction, and a double isolator in the transmitter modulator parallels the receiver modulator design. The dual isolator is omitted in the single-frequency carrier supply modulators.

In the carrier supply modulators, the low frequency is fed to the two diodes in parallel from a 75-ohm cable. Simple networks are used to match the cable impedance to the impedance of the two diodes. For the transmitter modulator, however, the IF signal is supplied to the diodes through a one-stage amplifier. The purpose of the amplifier is to isolate the 75-ohm cable from the diode IF impedance, which changes considerably over the 30-mc wide IF band. This amplifier employs a 448A which operates in the same manner as in the IF main amplifier. An input transformer matches the 75-ohm cable to the grid of the tube. A variable series inductor is in the plate circuit to correct for a slope in the RF output over each channel, and a relatively low plate load is used to minimize the gain-frequency variations due to variation of diode IF impedance.

With a microwave carrier (BO) input of +27 dbm and an IF input

of 0 dbm to the one-stage amplifier, the microwave output power of the transmitter modulator is +7 dbm. The output is flat to within 0.05 db over a 20-mc band and flat to within 0.25 db over a 32-mc band. The BO energy appearing at the output is at least 20 db below its power at the input to the modulator.

The 427A diode is capable of giving conversion gain if sufficient negative bias is applied to it, the gain increasing as the negative bias is increased. This is due to the fact that the 427A diode actually is a variable capacitance p-n junction, or varactor, and, therefore is low loss. This phenomenon was demonstrated first on the laboratory version of this diode by Uhlir.[7] However, the impedance of the 427A diode becomes more frequency sensitive as the gain is increased, and for this reason the diodes are operated at relatively low conversion gain in the TH equipment. A bias of -1 volt gives approximately 0-db gain, which meets system requirements for all high-level modulators except the 252-mc carrier supply modulator. The latter is operated at -4 volts bias to offset the loss of gain inherent in a higher frequency input. In all cases the required bias is obtained from a resistor in series with the diode.

III. TRAVELING-WAVE TUBE AMPLIFIER

3.1 *General*

The TWT microwave amplifier has a power output of five watts and uses the Western Electric Company 444A tube, which is the production version of a 6000-mc TWT described in an earlier paper.[8] A photograph of the amplifier is shown in Fig. 24, and a line drawing of a cross section through it in Fig. 25. The amplifier uses a permanent magnet focusing structure which consists of two permanent magnets, two pole pieces, a field straightener, and a movable gun shield. The magnets and the associated pole pieces provide a uniform magnetic field along the axis of the electron beam of the traveling-wave tube. The field straightener unit is placed concentrically with the axis of the electron beam between the input and output waveguide to reduce any transverse field present. The gun shield which is near the electron gun region of the tube controls the field at the cathode as well as providing a means for magnetic focusing of the electron beam through the use of mechanical controls. Waveguide circuits are used to couple signals into and out of the tube. The impedance match between the waveguides and the helix of the tube is adjusted by mechanical control of the positions of the waveguide along the axis of the tube and of a shorting plunger in the waveguide.

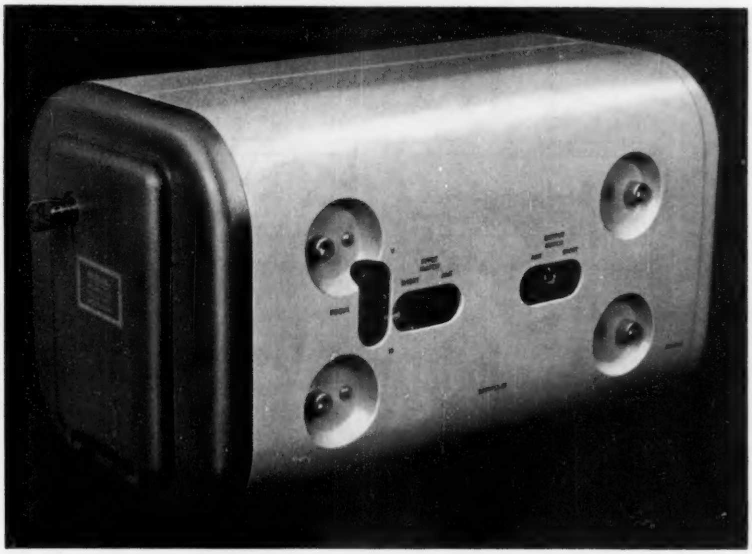


Fig. 24 — The TH traveling-wave tube amplifier.

The collector and the tube envelope are air cooled. The entire structure is enclosed in a magnetic shield to eliminate interaction with adjacent TWT amplifiers and other circuits. In addition, the shield and the associated mechanical and electrical interlocks protect operating personnel from the high voltages required for the tube.

The TWT amplifier has two applications in the transmitter; in the signal path (transmitter amplifier) and in the carrier supply path for the transmitter modulator (carrier supply amplifier). Requirements for the transmitter amplifier will be considered first.

The transmitter amplifier power output requirement is +37 dbm (5 watts). The allowable gain is dependent on the noise figure of the traveling-wave tube. The 444A has a noise figure of 27 db to 30 db, which is sufficiently high to make some contribution to the over-all repeater noise figure. The nominal gain of the repeater is 62 db (+37 dbm output, −25 dbm input), and the noise figure of the average-age radio receiver is about 11 db. If we allow the transmitter amplifier to make an additional contribution of 0.3 db to the repeater noise figure, then the average gain allowable for the amplifier is found to be about 33 db. The gain of the amplifier is therefore held to the range of 30 db to 35 db. In those cases where the amplifier gain exceeds 35 db at +37

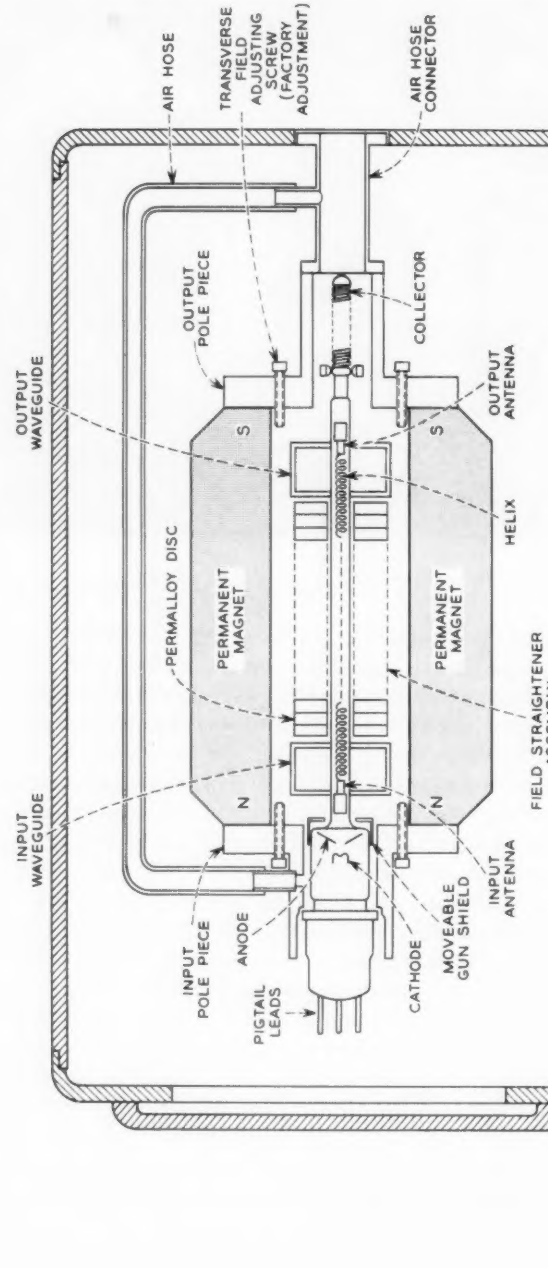


Fig. 25 — Cross section of traveling-wave tube amplifier.

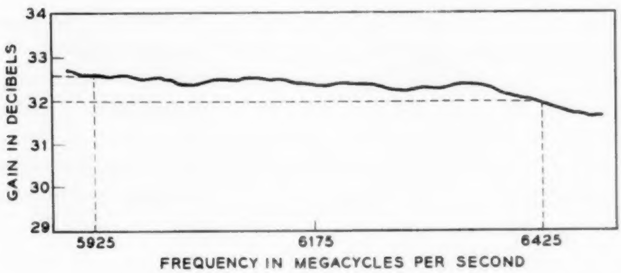


Fig. 26 — Typical gain-frequency response of the traveling-wave tube amplifier.

dbm output, the helix voltage is lowered until the gain is reduced to 35 db.

The gain-frequency response of the amplifier is substantially flat over the 5925 to 6425-mc band, and a typical example is shown in Fig. 26. The input return loss is 25 db minimum to provide a proper termination for the channel bandpass filter. The active or hot output return loss is about 10 db and is controlled principally by internal reflections in the tube. At least 25 db return loss is needed for terminating the channel separation network, and an isolator is used to meet this requirement.

Spurious radiation requirements call for second harmonic (12 kmc) output at 50 db below the carrier. As the amplifier at +37 dbm output may have the second harmonic down only 25 db, a low-pass filter providing a minimum of 25 db attenuation to the second harmonic is inserted immediately following the amplifier. Since 12-kmc energy can be propagated in WR 159 waveguide in several modes, it is not attenuated appreciably by the channel bandpass and channel separation networks.

The carrier supply amplifier output power requirement is +27 dbm (0.5 watt). Here, also, the allowable gain is determined by the noise figure of the 444A TWT. Noise on the carrier supply at BO frequency causes phase modulation, which appears as noise on the signal. To minimize this noise contribution, an input power of about +6 dbm is used, and the TWT gain is reduced by lowering the helix voltage. This provides a high carrier-to-noise ratio.

The structure associated with the TWT must focus the electron beam, couple energy in and out of the slow-wave structure, and dissipate the heat generated by the expended electron beam. Other design requirements arise from over-all system considerations. The economics of continuous operation in unattended remote repeater stations dictates that

the power required to focus the beam and to cool the tube be minimized. The compact arrangement of the radio equipment demands that the external magnetic fields of the focusing structure be negligible, or they will adversely affect the operation of nearby equipment. Furthermore, as the electron structure is the only part of the amplifier that deteriorates with age, its replacement in the field with a minimum number of simple adjustments is required for desirable operating and maintenance practices.

3.2 *Magnetic Circuit*

The description of the magnetic circuit divides conveniently into three parts. These correspond to the regions surrounding the electron gun, the helix, and the collector of the TWT. To a greater or lesser extent, the distribution of magnetic field in all three regions affects beam focusing, noise performance, and collector efficiency.

A minimum longitudinal focusing field of 580 oersteds is provided for the proper operation of the 444A tube. To achieve long life, the intercept current to the helix is kept to less than 2 per cent of the beam current of 40 ma. This requires that the maximum value of the transverse field not exceed about two oersteds.

The focusing field could be generated either by a solenoid or by permanent magnets in either a periodic or straight field configuration. The cost of providing electric power to operate the solenoid, however, is prohibitive. Although the periodic focusing method has the advantage of a size and weight reduction when compared to the straight-field method, this is not of primary importance in a ground installation such as the TH system. The straight-field structure is inherently simple in form and lends itself more easily than the periodic structure to the introduction of the broadband waveguide coupling circuits required to provide the necessary transmission performance. Shielding of the periodic structure, to control external fields, is relatively easy compared to the straight-field structure. However, at the time of this development, the high coercive force magnetic materials available for periodic structures were not sufficiently temperature insensitive to be practical. The structure chosen was therefore of the straight-field type.

The design of the magnetic circuit, shown in Fig. 25, can be achieved from the procedures described by M. S. Glass.[9] A prototype structure, having no external magnetic shield, required $19\frac{3}{4}$ pounds of Alnico VI to produce the required 580 oersteds in the 7.1-inch gap between the pole pieces. In the final design the addition of the external shield increased the magnet loading appreciably. As a result, it was not only

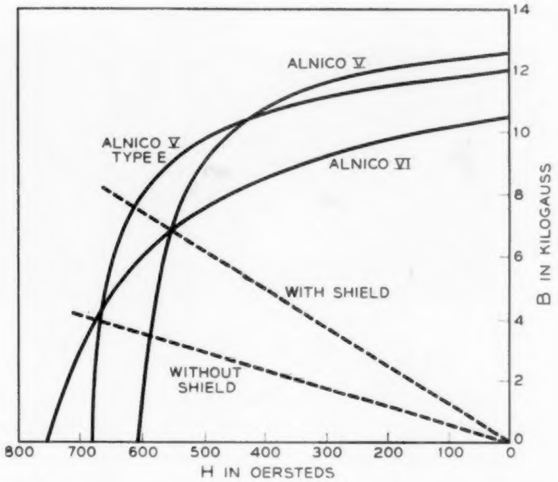


Fig. 27 — B-H curves showing effect of shielding on permanent magnet loading.

necessary to increase the amount of magnetic material to 27 pounds but to use Alnico V, type E, in place of Alnico VI. Fig. 27 shows the B-H curves for Alnico V, Alnico V type E, and Alnico VI, together with the load lines for the structure with and without shielding. It can be seen that the high coercive force of Alnico VI could not be utilized due to the loading of the shield. A plot of a typical longitudinal field distribution is shown in Fig. 28.

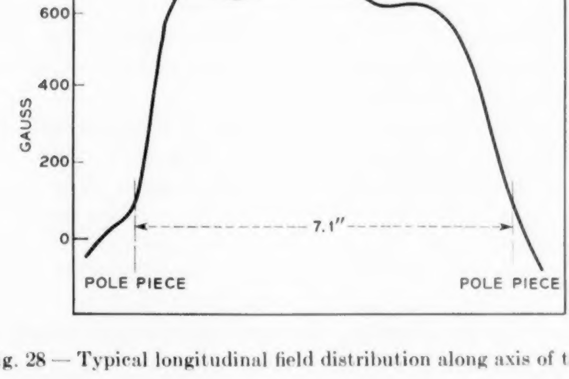


Fig. 28 — Typical longitudinal field distribution along axis of tube.

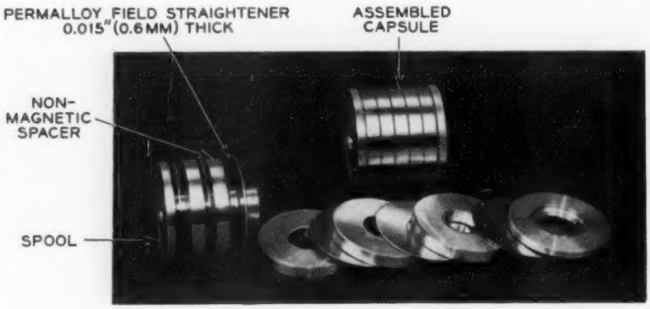


Fig. 29 — Field-straightener assembly.

To realize the transverse field requirements, it is necessary first to align the axis of the helix accurately with that of the magnetic circuit. If the net transverse component is to be less than two oersteds, the maximum mechanical misalignment must be less than 2/600 radian or about 0.2 degree. Secondly, it is necessary to provide as uniform a magnetic field as possible. Since the two magnets in the circuit are poled alike, minor differences between the magnets generate transverse fields. In the space between the waveguides these transverse fields are controlled by the field-straightener assembly, which is shown in Fig. 29. This consists of 18 thin Permalloy discs spaced $\frac{3}{16}$-inch apart. The high permeability of the discs normal to the axis of the tube shorts out the transverse fields. The discs are assembled in three identical cartridges, six discs to a cartridge, and are encased in a precision-machined aluminum casting. As shown in Fig. 30, the complete field-straightener assembly, in turn, is suspended between the waveguides by the two nonmagnetic stainless steel rods that connect the pole pieces. In the region occupied by the waveguides, field straighteners cannot be used, and the transverse fields in these regions are controlled by the adjustment of four iron screws inserted through each pole piece. Fig. 31 shows a typical transverse field plot.

Special attention was given to the shape of the field buildup at the ends of the structure, particularly at the cathode or input end. The 444A uses a converging Pierce-type electron gun, which is normally operated with the cathode completely shielded from the magnetic field.[8] Best focus results if the magnetic field is introduced in the region between the accelerating anode and the point where the electrostatic fields in the gun cause the beam to reach its minimum diameter in the absence of the magnetic field. However, the noise figure is importantly affected

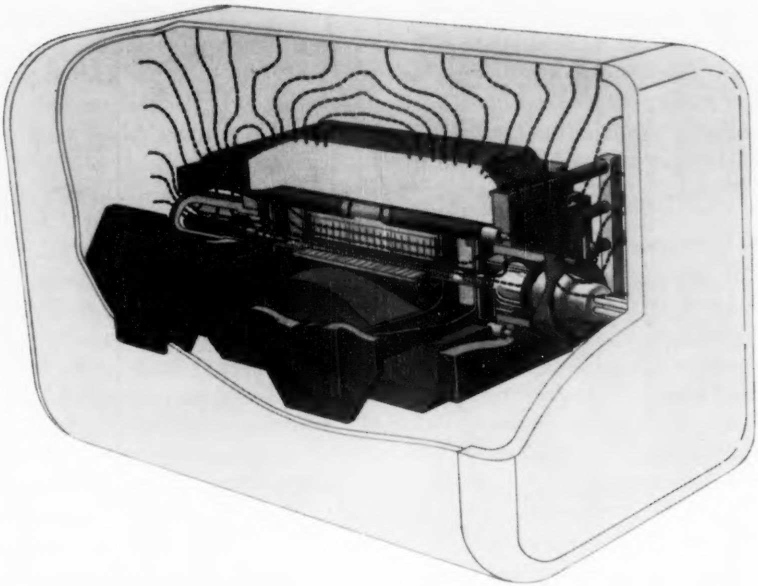


Fig. 30 — Magnetic assembly of the traveling-wave tube amplifier.

by the amount of focusing field in the gun region. Measurements made with a minimum field at the cathode showed that at the +37 dbm rated output power, the noise figure of the tube increased from about a low-power output value of 30 db to 40 db or higher. This is attributed to a growing noise wave on the electron stream, which depends on, among other things, the drive on the tube.[10] This growing noise wave can be

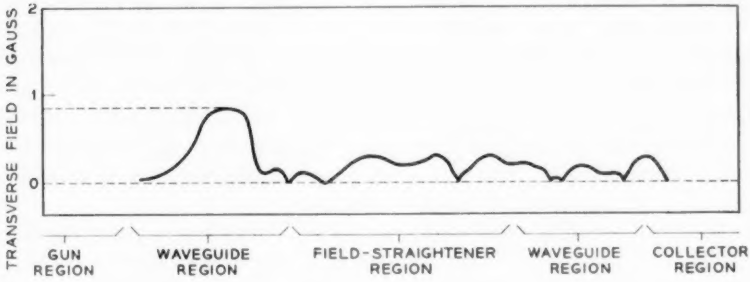


Fig. 31 — Typical plot of transverse field along axis of the tube.

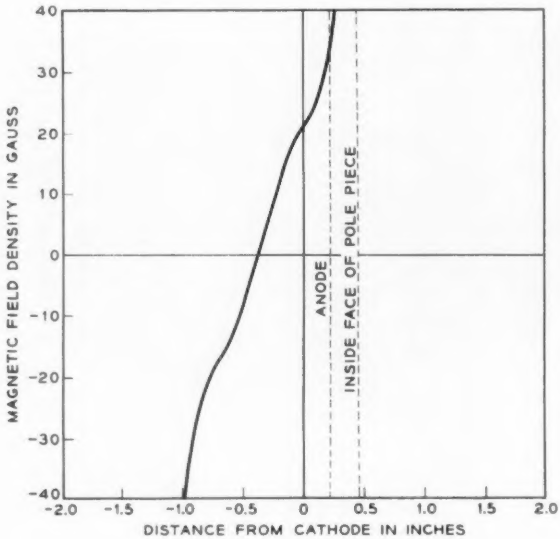


Fig. 32 — Longitudinal field in gun region of the traveling-wave tube amplifier.

reduced by increasing the magnetic field at the cathode. A field of 15 gauss is adequate to maintain the noise figure relatively constant throughout the tube operating range. The gun region magnetic structure is designed accordingly. In production, the cathode field is adjusted by selection of a suitable size soft-iron ring slipped over the gun-end shield.

As shown in Fig. 25, the magnetic field configuration in the gun region is controlled by the end shield and the gun shield, and by the magnetic material (Kovar) details used in the base of the tube itself. The most effective parameter for control of the field is the design of gun shield. The magnitude and shape of the field depend on the length of the gun shield, the size of the hole at the helix end of the shield, the type material and thickness of the gun shield face and walls, and the longitudinal positioning of the shield with respect to the tube and the rest of the magnetic circuit. Optimization of the dimensions provides the desired conditions as shown in Fig. 32.

Besides controlling the field conditions in the gun region, the gun shield is also used as a fine adjustment of the focus, by two controls which position it in a plane perpendicular to the tube axis. This movement produces a small transverse magnetic field in the vicinity of the hole in the gun shield but causes essentially no change in the field con-

ditions elsewhere. This small transverse field is sufficient to provide a successful focusing adjustment with the close mechanical tolerances imposed on the 444A and the magnetic circuit.

The magnetic field conditions at the collector end of the circuit are not particularly critical. Excessive helix intercept current results unless the focusing field is maintained close to a value of 600 oersteds out to the end of the helix, but the rate of field decay beyond the end of the helix is not important.

The design of the external shield is controlled by the tolerance of adjacent equipment to magnetic fields. To keep the size reasonable, the shield must be located in a region of high field intensity. Under these conditions, the magnetic properties of low carbon steel are superior to those of most alloys normally used for shielding. The sides of the shield are fabricated from $\frac{1}{8}$-inch low carbon steel. The thickness is determined by the allowable external field. Magnetic iron castings are used at the ends to complete the shield. The external field is less than one oersted at one-half inch from the center of the sides of the complete assembly.

A large dc-powered solenoid, having an inside diameter of $12\frac{1}{4}$ inches and a length of 24 inches with a magnetic field intensity of 4000 oersteds, is used to magnetize the complete assembly.

Specialized flux-measuring equipment is required for determining the longitudinal and transverse components of the focusing field. The helix of the 444A has an inside diameter of 0.080 inch. To measure the field in the region of the electron beam, longitudinal and transverse search coils of approximately the same diameter as the helix are used. These coils mount in an accurately machined rod which is supported from the surfaces used to locate the TWT in the magnetic structure.

Two types of search coils and associated equipment are used. The first type, used primarily for longitudinal field measurement, consists of an air-core coil and an integrator of a type similar to that described by Cioffi.[11] The second type, primarily for transverse-field measurement, uses a permalloy-core coil as a magnettor with its associated equipment.[12] In both instruments the field is plotted on an X-Y recorder as a function of search coil position.

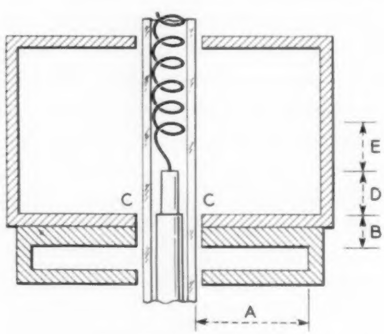
3.3 *Helix-to-Waveguide Transducer*

The microwave signal is coupled into the TWT by surrounding an antenna attached to the helix with a section of standard rectangular waveguide, to form a transducer. As shown in Fig. 33, a radial-line choke is incorporated to form an effective ground plane at the end of the helix. Dimensions A and B are both a quarter-wavelength long, and a micro-

wave short circuit appears across the opening CC. The performance of the choke is broadbanded by making the characteristic impedance of section A high relative to that of section B. Ordinarily, many turns of the helix are exposed to the field in the guide, which results in weak coupling. In this transducer the coupling is increased by stretching out the last turn of the helix to aid in matching it to an antenna or post which can be coupled to the waveguide. In Fig. 33 the height of the post is denoted by D and length of the last turn on the helix by E.

To a first approximation, the equivalent circuit of the helix-waveguide junction, as shown in Fig. 33, may be represented by the parallel combination of a conductance G and a susceptance jB connected across a transmission line; Y_0 is the characteristic admittance of the waveguide. For values of G near Y_0, the conductance is determined mainly by the height of the post and the susceptance by the pitch of the last helix turn. Increasing the post height increases G; increasing the length of the last helix turn decreases jB. By proper choice of dimensions, therefore, it is possible to obtain a match between helix and waveguide.

For small values of jB, G is almost constant with frequency, while jB itself decreases somewhat as the frequency is increased. If the waveguide is terminated by a microwave short circuit somewhat less than a quarter wavelength beyond the helix, this termination will look like a small inductive susceptance at the helix. This susceptance also decreases as the frequency increases. By dimensioning the post so that the value of G equals Y_0, and the last helix turn so that its capacitive susceptance, jB, cancels the inductive susceptance due to the short circuit, it is possible to produce a broadband match between helix and waveguide. Fig. 34 shows the return loss of the typical transducer over the 6-kmc band.



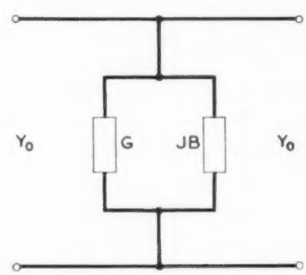


Fig. 33 — Cross section of helix-to-waveguide transducer and equivalent circuit.

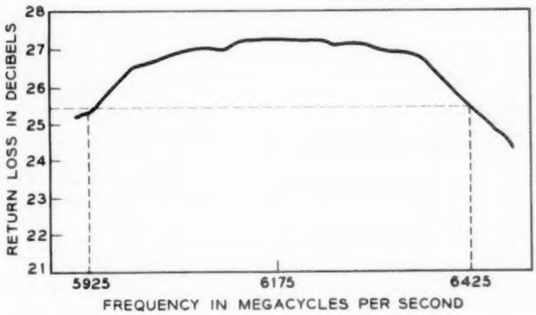


Fig. 34 — Return loss of typical transducer.

The same transducer arrangement is used at both ends of the helix. However, while the input match of the amplifier is that of the helix-waveguide transducer, the output match is controlled by minute periodic imperfections in the helix.[8] Any reflection of the output signal from the circuits beyond travels backward along the helix, essentially undiminished, to the helix attenuator. Impedance variations along the helix resulting from mechanical imperfections or at the helix attenuator cause a portion of the reflected signal to be re-reflected toward the output and be amplified by the gain existing between the point of reflection and the output. The output transducer connects to a microwave harmonic filter whose out-of-band impedance is reactive. Over a wide range of frequencies, therefore, the helix is mismatched either by the filter or the transducer. As a result, the traveling-wave tube itself must be short-circuit stable if the amplifier is not to oscillate.

In addition, it is necessary to control any external coupling between output and input. The radial line chokes do not provide perfect ground planes at all frequencies; some energy escapes along the extension of the helix post, which can be coupled to the other end of the structure through a crude cavity formed by the magnetic shield. To reduce the leakage, beryllium copper fingers between the outside of the choke and the pole piece are provided. The high-voltage connection from the terminal board at the input end to the collector is a second source of coupling. Lossy beads are put on the collector lead to provide attenuation in this path.

3.4 *Collector Cooling*

In normal operation, the beam current of the 444A is 40 ma and the collector voltage is 1250 volts, providing 50 watts to be dissipated in

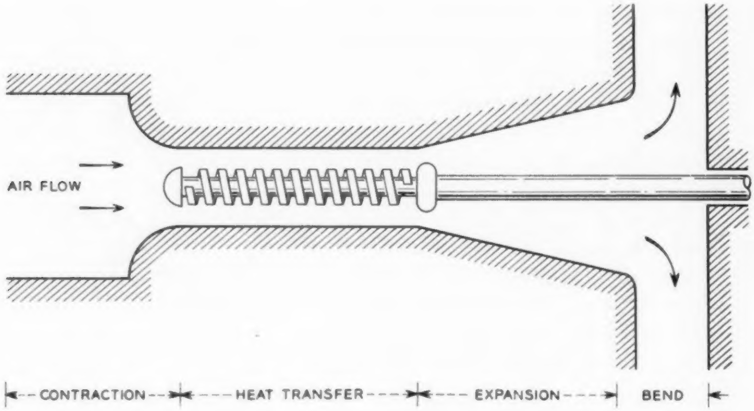


Fig. 35 — Cross section of the collector cooling chamber.

the collector. Since the collector is only about $1\frac{3}{4}$ inches long and $\frac{5}{16}$ inch in diameter, the heat dissipation is approximately 30 watts/in^2. To prevent serious outgassing, the temperature of the collector must not exceed 300°F, and forced air cooling is employed.

To minimize the pumping power required for forced air cooling, it is necessary to minimize the product of pressure drop and volume rate of flow. Fig. 35 shows a cross section of the collector cooling chamber. For a given rate of flow, the energy losses in each section are directly proportional to the corresponding pressure drops, but only the energy loss in the heat transfer section contributes to cooling the collector. The expansion and compression sections were designed for minimum pressure drops.

The pressure drop across the heat transfer section and the heat transfer to the cooling air are both a function of the air velocity and the roughness of the collector surface. The collector surface conditions that would minimize the pumping power for a given heat transfer were determined experimentally. With an initial air temperature of 70°F, the required static pressure drop and volume rate of flow are $4\frac{1}{4}$ inches of water and four cubic feet per minute. This corresponds to a pumping power of approximately two watts.

A small quantity of air is piped from the collector inlet to the electron gun envelope to keep its temperature from exceeding 250°F. The final pressure and volume requirements for the amplifier, which allow for a higher ambient temperature, are eight inches of water and six cubic feet per minute. The corresponding pump power is approximately $5\frac{1}{2}$ watts.

3.5 *Thermal and Ion Oscillation Noise*

The contribution of the TWT amplifier noise figure to the repeater performance has been discussed. To measure the TWT frequency modulation noise, a microwave signal is applied to the amplifier from a klystron. The carrier and noise sidebands appearing at the amplifier output are amplified and demodulated by a skeletonized radio receiver and FM receiver. The recovered video noise spectrum, of the familiar triangular shape, can be measured with an ordinary selective analyzer or AM detector. The measuring circuit is calibrated by using a known noise source.

A second cause of noise impairment in the system due to the TWT results from ion oscillation in the electron stream, which gives rise to spurious modulation of the carrier. In the 444A this modulation appears as a relatively narrow band of energy in excess of normal tube noise, usually at about 2.8 mc each side of the signal carrier. By using a relatively short period of aging under full RF drive conditions during manufacture, it has been possible to hold the ion oscillation noise at baseband to 10 db or less above normal tube noise. As the ion density within the tube decreases with operation, the excess noise disappears after a few hundred hours.

The allowance of 10 db above thermal noise for ion oscillation assumes that only a few new tubes would be installed in a 4000-mile system at any time within the few hundred hours required for tube ion cleanup. Since the thermal noise contribution of one repeater is 21 db below that of the system, and the TWT contribution 12 db below that, the increase in noise due to one tube with ion noise 10 db above thermal in one repeater is negligible. In fact, if 10 tubes were changed at once, the increase in noise for a 4000-mile circuit would be only about 0.3 db, which would appear in only a few dozen telephone channels for a few hundred hours.

3.6 *Mechanical Design*

The over-all dimensions of the amplifier are approximately $9\frac{3}{4}$ by 10 by $17\frac{1}{4}$ inches, and the complete unit weighs 86 pounds, of which the permanent magnets account for 27 pounds and the external shield accounts for 36 pounds. The front and rear of the shield are cast iron and engage interlocking tabs on the ends of the side covers. Large aluminum castings which attach to each pole piece suspend the magnetic circuit inside the shield. Four screws, accessible only through the front door, run the full length of the amplifier and secure the front and rear castings. The amplifier mounts on a vertical panel on four large brass pins that protrude from the panel. These four pins engage the aluminum

castings through holes on the side cover. The ends of the mounting pins are tapped to accept captive screws which come through from the opposite side cover. A locked end door gives access for replacement of the electron tube. To obtain the key to unlock the door, it is necessary to lock the high voltage power supply to the OFF condition.

IV. EQUALIZATION

Equalization is an area which is still under development, as discussed in Ref. 1. The description here is confined to the non-adjustable equalizers mounted on the broadband receivers. One type is the basic gain and delay equalizer used in every receiver. It is intended to compensate for departures from ideal of the transmission characteristics of one complete radio section. Its characteristic will eventually be determined by the average transmission characteristics of a large number of radio sections. Fig. 36 shows the average characteristic of the first 30 manufactured transmitter-receiver pairs, as determined from factory point-by-point measurements. The envelope delay distortion (EDD) curve includes the limiter, but this was necessarily excluded in determining the loss characteristic.

The present basic equalizer is made up of five delay sections in tandem with a gain-correcting section. The delay sections are 360° all-pass networks in a bridged-T, constant-R configuration. The gain sec-

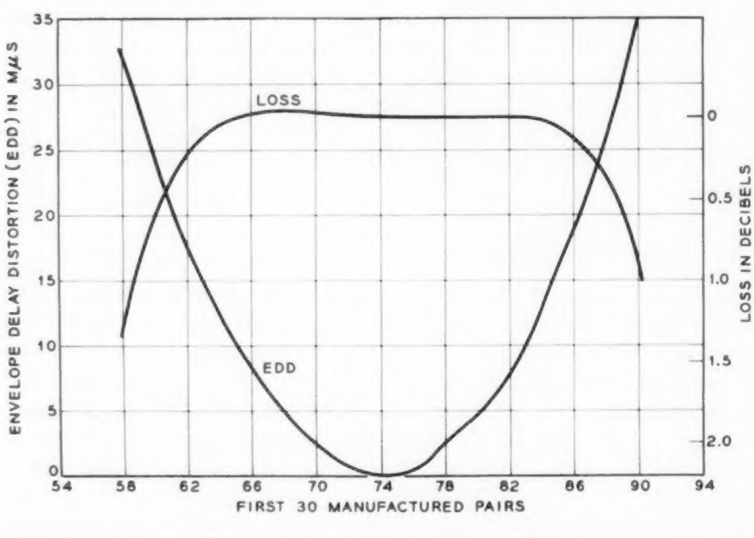


Fig. 36 — Average characteristic of the first 30 manufactured transmitter-receiver pairs.

tion is similar. Each section is individually shielded in cans which are attached to a common U-shaped chassis. The equalizer matches the repeater to about ± 1 mμs over 60 mc to 88 mc.

The transmission characteristics vary from channel to channel, for several reasons, although much effort is taken to minimize this. Part of the variation is due to the differences in frequency, even though the microwave filters are specifically designed for constant bandwidth. Part is due to unavoidable manufacturing variations. Part arises from the varying number of channel separation networks which are traversed on the path to the antenna. The most important part of this channel-to-channel variation is linear envelope delay distortion, or delay slope. Accordingly, two sizes of delay slope equalizer are made available: $+1$ mμs/mc and -1 mμs/mc. Over-all EDD measurements of a switching section determine the amount of slope correction needed for each channel. The necessary number of slope equalizers are then distributed among the repeaters of the channel. For the worst channel, one-third of the repeaters may need delay slopers. They are mounted on the receiver, as indicated in Fig. 1. Their design is very similar to the basic equalizer: two delay and one loss sections are used. The delay shapes are linear to about ± 0.2 mμs over 60 mc to 88 mc.

V. ACKNOWLEDGMENTS

A project of this magnitude represents the work of many competent development engineers, and this article could not have been written without their assistance. Special mention should be made of D. R. Jordan, P. R. Wickliffe and the late F. W. Koller for their work on the traveling-wave tube amplifier. H. W. Andrews, M. G. Davis, and P. I. Sandsmark contributed to the modulator designs, and A. J. Giger was largely responsible for the amplifier-limiter and carrier resupply designs. Many others are to be thanked for their contributions.

APPENDIX

To show the dependence of AM/PM conversion on circuit parameters.

The phase shift, θ, between the driving voltage V_L and the current I_p across a parallel RLC circuit for small frequency deviations Δf from center frequency is

$$\theta = -2Q \frac{\Delta f}{f_0}, \tag{6}$$

where f_0 is the resonant frequency, and

$$Q = 2\pi f_0 RC. \tag{7}$$

In a clipping network, like Fig. 15, but with perfect diodes, the funda-

mental component V_L of the output voltage is constant in amplitude. At f_0 the apparent parallel resistance R_A is then given by

$$R_A = V_L/I_p. \tag{8}$$

Now if we assume that (at $\pm V_{co}$) the diodes are clipping the incident sine waves symmetrically and so close to the base line that the voltage across the diodes appears essentially as a square wave, then the fundamental component of this square wave is given by

$$V_L = \frac{4}{\pi} V_{co}. \tag{9}$$

Combining (6) to (9) we obtain

$$\theta = -16 \frac{\Delta f c V_{co}}{I_p}. \tag{10}$$

Now AM/PM conversion is defined by

$$P = \frac{d\theta}{dV_i/V_i} \tag{11}$$

where $d\theta$ is the infinitesimal change of phase in radians caused by a fractional change dV_i/V_i of input voltage. Since I_p is proportional to V_i, (11) may be written as

$$P = \frac{d\theta}{dI_p/I_p}. \tag{12}$$

P may be expressed in degrees per db by multiplying by the factor 6.6. Differentiating θ with respect to I_p in (10) and substituting in (12) we obtain

$$P = 16 \frac{\Delta f c V_{co}}{I_p}, \tag{13}$$

or if P is expressed in degrees per db,

$$P = 105.6 \frac{\Delta f c V_{co}}{I_p}. \tag{14}$$

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The TH Radio Microwave Carrier Supply System

By C. P. SUSEN and J. J. DEGAN

(Manuscript received June 12, 1961)

The TH microwave carrier supply system furnishes all of the carrier frequencies required at a radio repeater station. All the frequencies are harmonically derived from a common, highly stable, 14.82593-mc crystal oscillator, to avoid interference problems. Total reliance on a common generator of carriers for an entire station calls for a highly reliable and adequately protected carrier supply. To this end, a complete standby supply is arranged with fast automatic switching to replace the regular supply when needed. Special emphasis is placed on conservative circuit design, careful selection of components, adequate monitoring and trouble detection facilities, and provisions for routine maintenance.

I. INTRODUCTION

As described briefly in a preceding paper,[1] the common microwave carrier supply system supplies all the carrier frequencies (often called beat oscillator, or BO, frequencies) needed for a fully equipped repeater station. The major portion of the system is in three sliding-rack bays[2] which house regular and standby frequency generators, together with switching and control circuits. This equipment supplies four harmonically related frequencies (29.7 mc, 59.3 mc, 6301 mc and 6049 mc). Other parts of the carrier supply system are the carrier distribution networks which carry these frequencies as needed to the various transmitters and receivers, where carrier supply modulators[3] produce the specific frequencies desired.

Frequency stability and over-all reliability are prime objectives for the microwave carrier supply. Frequency stability is given by a crystal-controlled, thermistor-compensated oscillator with a long term stability of better than 10 ppm over operating temperatures of 40°F to 140°F. Reliability is obtained by the regular-standby arrangement. This requires various detectors, control circuits, and switches of a reliable nature to identify a failure quickly and replace the failed regular carrier

supply with the operative standby supply. A failed regular supply is replaced by the working standby supply in about five milliseconds. Failures of the standby supply are alarmed; no switching results for this case. Various manually controlled switches are provided for maintenance work.

All of the power connections and test points are independent to prevent simultaneous failure of the regular and the standby supplies. Also, to the extent possible, the equipment layout is arranged to minimize accidental contact by maintenance personnel with a carrier supply which is actively connected to the distribution circuit.

II. BLOCK DIAGRAM

A simplified block diagram of the microwave carrier supply is shown in Fig. 1. The 14.8-mc signal from either the regular or the standby crystal oscillator feeds both the regular and the standby frequency generators through the low-frequency switching circuit. Both generators are supplied by the same oscillator to prevent beats which might be caused by a minute frequency difference between the two oscillators. This eases

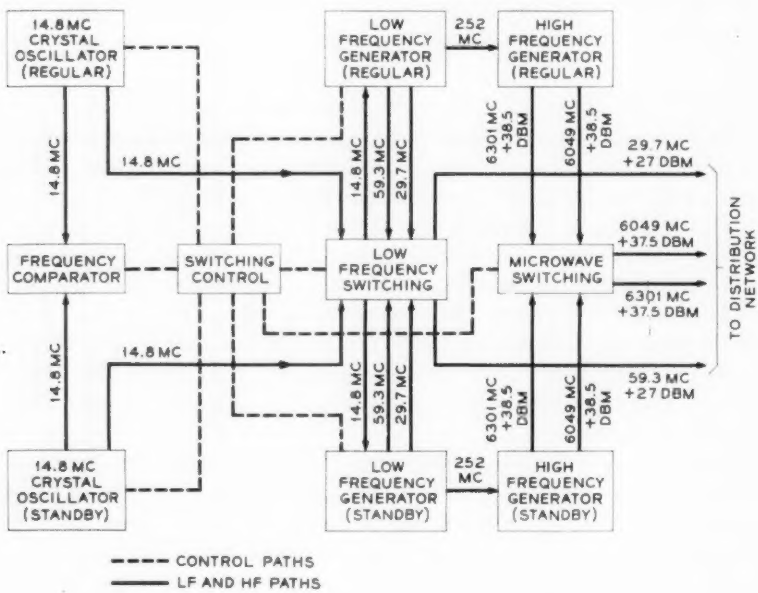

Fig. 1 — Simplified block diagram of the microwave carrier supply.

the isolation requirements on all the switches except those associated with the 14.8 mc. The low-frequency generator supplies the second harmonic (29.7 mc) and the fourth harmonic (59.3 mc) of 14.8 mc by successive frequency doubler circuits. A third output is the 17th harmonic (252 mc) which is supplied to the high-frequency generator. Here it is multiplied 24 times in three doublers and a tripler (cavity type multipliers). The resulting microwave frequency is the 408th harmonic (6049 mc). This in turn is added to the 17th harmonic to obtain the 425th harmonic (6301 mc) of the crystal oscillator frequency. Amplifiers are provided to obtain the required output powers of +27 dbm at 29.7 mc and 59.3 mc and of +37.5 dbm at 6049 mc and 6301 mc. The switching system normally connects the outputs of the regular frequency generators to the distribution system; if a failure or serious degradation of the regular equipment is detected, an automatic switch transfers operation completely to the standby equipment.

Of the three sliding-rack bays, the bay on the left contains the regular frequency generators and an identical bay on the right contains the standby generators. The middle bay contains the regular and standby 14.8-mc oscillators, the low-frequency switching circuits, and the switching control circuits. The microwave switching equipment is mounted across the top of the three bays. Regular and standby equipments are fed from independent ac power sources. Solid-state rectifiers are mounted on the racks to obtain the dc voltages for the individual circuits on that sliding rack.

2.1 *Distribution*

The carriers at 29.7 mc and 59.3 mc are each connected through coaxial cable to a narrow band-pass filter which suppresses other harmonics of 14.8 mc and then to a hybrid transformer tree. Each signal is divided four times to provide a total of 16 equal outputs at 29.7 mc and 12 equal outputs at 59.3 mc. These 28 outputs are connected through coaxial cable to the various broadband transmitters and receivers and to the auxiliary channel receiver circuits. The four auxiliary channel transmitters do not use either the 29.7-mc or the 59.3-mc carrier.[4] With a nominal input power of +27 dbm, the nominal power at the outputs is +13 dbm. This assumes 0.5-db insertion loss for each transformer and the 3-db loss due to equal power division in each transformer.

The carriers at 6049 mc and 6301 mc are subdivided by trees of waveguide power splitting junctions, as shown in Fig. 2. These junctions provide an equal division of power between the two output ports 2 and 3, when a signal is connected to the input port, 1. Each microwave signal

is divided to provide a total of 8 equal outputs for broadband transmitter bays, 4 equal outputs for broadband receiver bays, and 8 equal outputs for the auxiliary channel transmitters and receivers. The outputs to the broadband channels are further split locally, once on the transmitter bays and twice on the receiver bays, to furnish a total of 32 equal outputs for the broadband transmitters and receivers. Connection to the broadband channel bays is made through waveguide and to the auxiliary channel bays through coaxial cable. Alternate stations on a TH route interchange 6301 mc and 6049 mc between transmitters and receivers as shown by the X and Y alternatives of Fig. 2.

With a nominal input power to the distribution system at 6049 mc and 6301 mc of +37.5 dbm, the nominal output power delivered to a

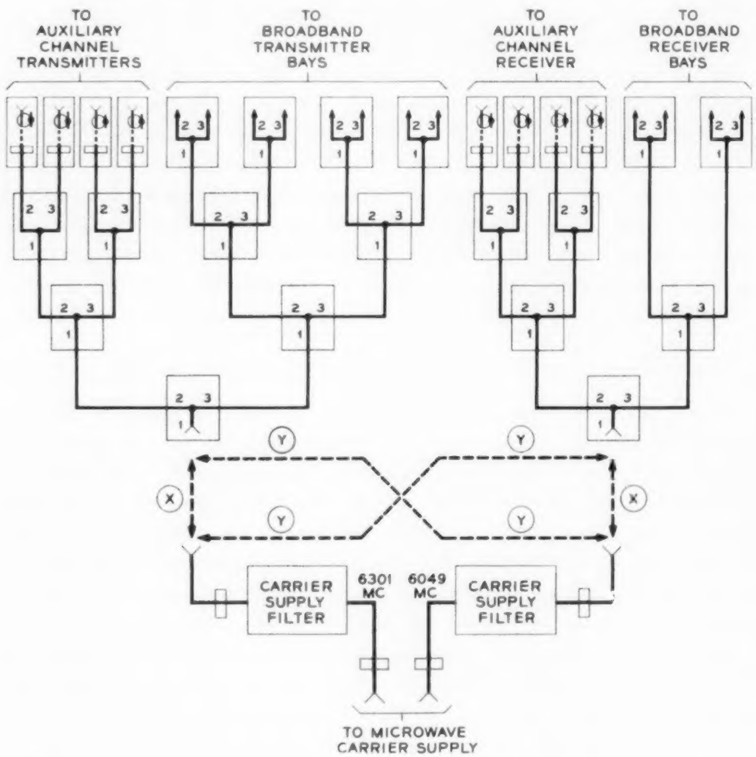

Fig. 2 — Block diagram of distribution circuits for the high-frequency microwave carrier supply.

broadband transmitter or receiver is +20 dbm and the nominal output power for the auxiliary channel equipment is +26.3 dbm. This assumes 0.2-db insertion loss for the junctions, 1.5-db insertion loss for the band-pass filters, and 0.1-db insertion loss for the transducers plus the 3-db loss due to the equal power division of the junctions.

III. 14.8-MC OSCILLATOR

This circuit, shown in block form in Fig. 3, uses two 404A pentodes, the first as a crystal-controlled oscillator and the second as a buffer amplifier. The modified Pierce oscillator circuit operates the crystal on its third overtone. Temperature compensation of the crystal may be provided, depending on temperature testing of the individual crystal. This compensation keeps the operating frequency within limits of ±148 cycles over a temperature range of 10° to 60°C. A variable inductor in the plate circuit is used as a vernier frequency adjustment to set the output within a few cycles of a standard frequency source of 14.82593

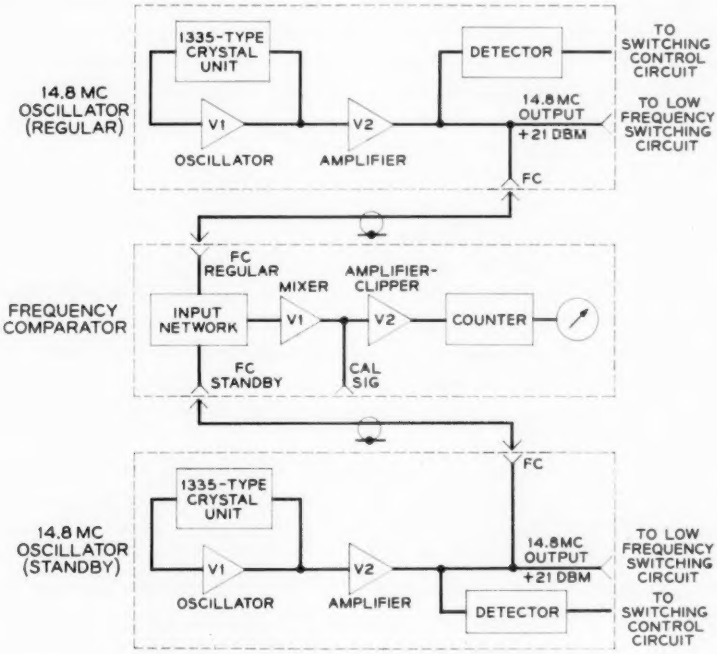

Fig. 3 — Block diagram of the 14.8-mc supply circuit.

mc. This inductor also prevents the crystal from oscillating at its fundamental frequency.

The second stage (V_2) is a class B amplifier with its output circuit tuned by a variable capacitor to the input frequency. The output coupling network is essentially a pi-type matching network between the plate circuit of V_2 and a 75-ohm output jack. Portions of the output are capacitively coupled to the frequency comparator and to a peak detector circuit. The latter furnishes a dc control signal to the switching control circuit. The coupling capacitor associated with the detector circuit is variable to allow the dc output of the detector to be set at the required level.

The output signal power is adjustable over a range of approximately 9 db by means of a potentiometer in the cathode circuit of the amplifier stage. Nominal output is adjusted to +21 dbm.

IV. FREQUENCY COMPARATOR CIRCUIT*

The two 14.8-mc oscillators are in continuous operation, and by comparing these two frequencies, relative changes in frequency can be detected. The frequency comparator does this and also provides a means of comparing either oscillator to a standard frequency. Both aural and visual monitors of the difference frequency are provided, and an alarm is issued if the difference exceeds 150 cps.

The building blocks of the frequency comparator are the input network, mixer, amplifier-clipper, and counter circuit as shown in Fig. 3. A simplified schematic is shown in Fig. 4.

The position of the frequency comparator is literally between the two 14.8-mc oscillators in both the mechanical and electrical sense. Thus, it could provide a cross coupling path between the two oscillators. To obtain the required coupling loss, a hybrid transformer is used which provides 40 db of isolation, and a 20-db pad is in series with each input signal. This provides a total of approximately 80-db loss between the oscillators.

The two input signals are combined in the hybrid transformer and are applied to the grid of V_1, a 404A pentode, producing the sum and difference frequencies in its plate circuit. The plate load of V_1 has a low-pass filter characteristic so that only the difference frequency appears at the input of $V_{2\text{-}A}$. This is half of a 396A dual triode operating as a class A voltage amplifier. The amplified sinusoidal signal is coupled to $V_{2\text{-}B}$, where it is converted into a rectangular wave by combined grid-

* This section was written by A. F. Perks.

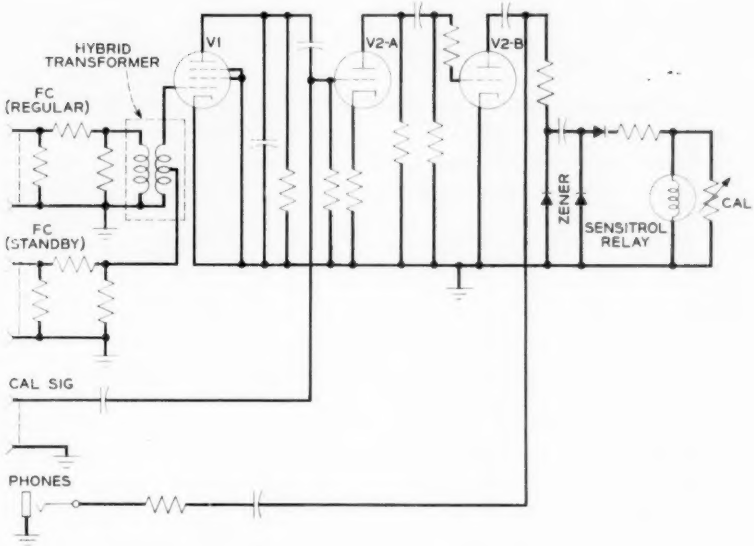

Fig. 4 — Simplified schematic of the frequency comparator.

circuit and plate current cutoff clipping. A voltage reference diode in the plate circuit of V2-B gives additional clipping and reduces any variations in voltage level due to tube aging.

The counter circuit converts the rectangular signal from V2-B into a train of rectified current pulses whose shape is independent of the frequency as long as the time constant of the counter circuit is much smaller than the period of the signal. The pulse train is used to deflect a 50-microampere Sensitrol relay. The average current through the winding, therefore, increases linearly with the repetition rate of the incoming rectangular signal.

The counter circuit is calibrated by inserting a 60-cps signal at CAL SIG and adjusting the potentiometer for a meter reading of 20 μa. A change of 1 μa in the meter reading will then correspond to a change of 3 cps in the difference frequency. Thus, for a difference frequency of 150 cps, the current will be 50 μa, at which time the Sensitrol relay contacts will make and an alarm will be initiated.

V. LOW-FREQUENCY GENERATOR

A block diagram of this circuit is shown in Fig. 5. With the exception of the last three stages, V8-V10, it uses 404A pentodes. The 14.8-mc in-

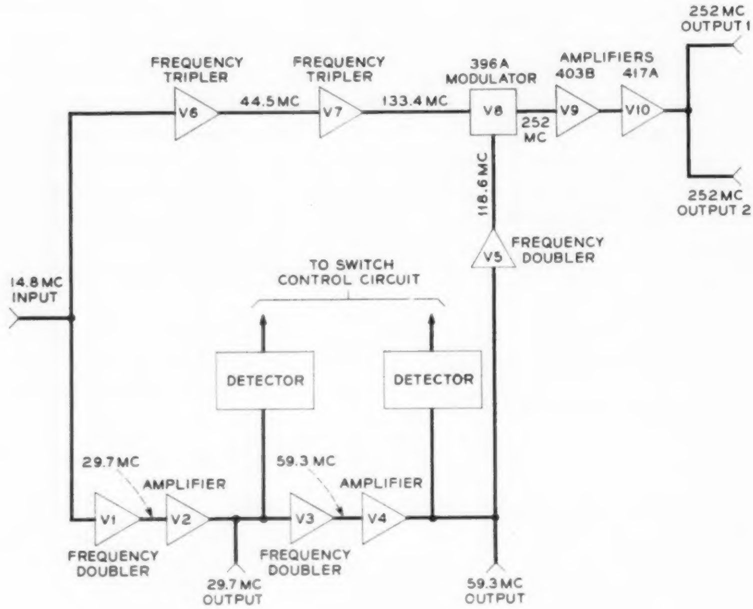

Fig. 5 — Block diagram of the low-frequency generator.

put signal at approximately +17 dbm is connected to two circuits, one a chain of three frequency doublers and two amplifier stages, and the other a chain of two frequency tripler stages. The outputs of the first and second doublers are amplified to provide the signals at 29.7 mc and 59.3 mc. The outputs of the third doubler and second tripler are fed to a balanced modulator whose output is tuned to 252 mc. The 252-mc signal is further amplified to provide the third output signal of the low-frequency generator.

For the most part, the circuits are fairly conventional. The power amplifiers, V2 and V4, provide a maximum output into a 75-ohm load of +31 dbm. Input signals for these stages are coupled from the preceding doublers through diode limiters, which maintain a constant input level and hence a constant output power for relatively large variations in signal level at the plate of the preceding stage. The output power is adjustable over a range of approximately 8 db with a nominal value of +28.5 dbm. A portion of the output is capacitively coupled to a peak detector circuit which furnishes a dc control signal to the switching control circuit. The coupling capacitor is variable to allow the dc output of the detector to be set at the required level.

The modulator stage (V8) uses a 396A twin-triode in a balanced modulator type circuit. The balanced secondary of the input transformer is tuned by a variable split-stator capacitor to the output frequency of V5 (118.6 mc). An analysis of the modulator, assuming a square-law characteristic, shows that the output circuit contains only sum and difference products. Even harmonics of both input signals tend to be cancelled by about 20 db, depending on the degree of balance of the two sections of the tube and on the balance in the input and output transformers. Both the balanced primary and the secondary of the output transformer are tuned to the sum frequency (252 mc). The difference frequency is far enough removed that no special filtering is needed. The secondary is capacitively coupled to the grid of the following stage.

A 403B pentode (V9) is used as an amplifier between the modulator stage and the final amplifier. Most of the tube bias voltage is developed across the grid resistor. A small cathode resistance provides the remainder of the bias and limits the plate current under zero signal condition to a safe value. This method of obtaining tube bias provides an output signal that is relatively independent of variations in the input signal level over a moderate range.

The final amplifier (V10) uses a 417A triode in a grounded-grid circuit. Signal from V9 is capacitively coupled to the cathode circuit. The output, nominally +22 dbm into 75 ohms, supplies the 252-mc shift modulator in the HF generator. A portion of this output is capacitively coupled to a 50-ohm coaxial line, to feed the high-frequency multiplier chain. Both outputs are controlled by a variable resistor in the cathode circuit of V10. The output at the 50-ohm jack may be adjusted independently over a limited range by a variable capacitor.

VI. HIGH-FREQUENCY GENERATOR

A block diagram is shown in Fig. 6. The 252-mc input passes through a narrow band (about 500 kc wide) cavity-type filter to a chain of harmonic generators, HG1 to HG4. Fig. 7 is a photograph of these. All harmonic generators are frequency multipliers using a 416-type planar triode mounted in a grounded-grid circuit.

The input of HG1 is fixed tuned. The output is a coaxial section of fixed length and is tuned by an adjustable screw plunger which acts as a variable capacitance between the plate of V1 and ground.

The coaxial output of HG1 is connected to the input of HG2, which is fixed tuned. The coaxial output section is tuned to 1512 mc by a movable quarterwave coaxial transformer. The combination of this section and two fixed quarter-wave sections provides an impedance transformation between the plate of V2 and the coaxial output cable.

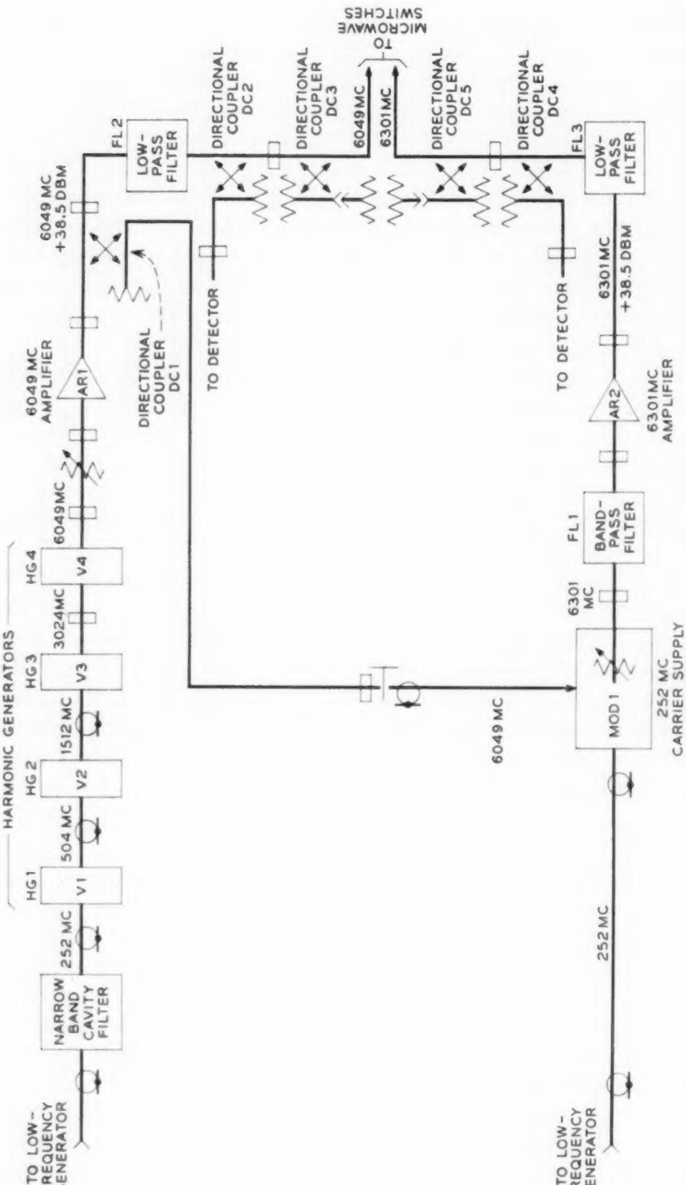

Fig. 6 — Block diagram of the high-frequency generator.

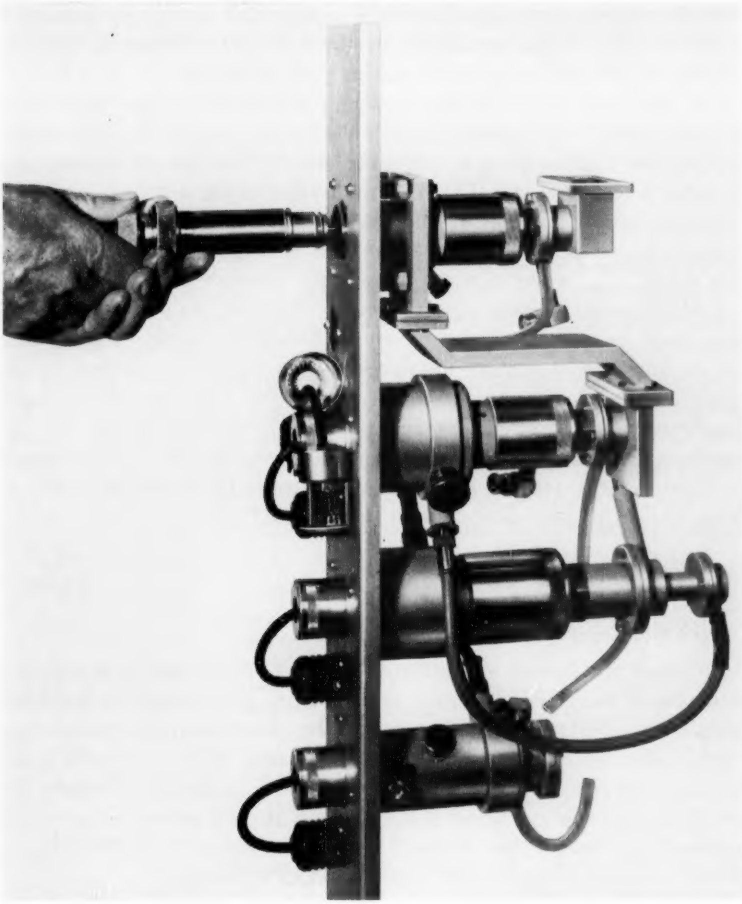

Fig. 7 — Chain of harmonic generators used in the high-frequency generator.

The coaxial input of HG3 is tuned to 1512 mc by an adjustable screw plunger which acts as a variable capacitance between the inner and outer conductors of the coaxial input section. The output section is coaxially tuned as in HG2 to 3024 mc. However, the output is waveguide-coupled to the last harmonic generator (HG4) through a coaxial-to-waveguide transducer, which is an integral part of the cavity structure.

The input signal to HG4 at 3024 mc is coupled to the cathode by an iris and a cylindrical cavity which surrounds the grid and cathode termi-

nals. The input circuit is tuned by an adjustable screw plunger which varies the susceptance across the cavity walls. The output is tuned to 6049 mc in a manner similar to the output tuning for HG2.

The output of HG4 is coupled through waveguide to the input of a traveling-wave tube amplifier, AR1. The main output at +38.5 dbm goes to the high-frequency switching circuit. Directional coupler DC1 furnishes a low-level signal at 6049 mc as one of the inputs to the 252-mc carrier supply modulator. The 6049-mc and 252-mc signals are mixed in the balanced modulator, and the output is connected to the input of a second traveling-wave tube amplifier, AR2, through a band-pass filter, FL1, which passes only the sum frequency. The 6301 mc output at +38.5 dbm goes on to the high-frequency switching circuit. Filters FL2, FL3 are low pass, to remove undesired second harmonics present in the amplified signal. Directional couplers, DC2 and DC4, furnish low-level signals to detector circuits which provide dc control signals for the switching control circuit. Couplers DC3 and DC5 furnish 0-dbm signals for measurement purposes. The traveling-wave tubes are identical to those used in the broadband radio transmitter.[3]

VII. SWITCHING AND CONTROL

7.1 *Reliability*

Although the largest single factor assuring carrier supply continuity is the use of the regular-standby arrangement, great attention has been given to reliability in the detail design, by conservative circuit design, careful selection of components, and avoidance of devices with questionable electrical and mechanical performance. Examples include the use of long-life electron tubes at low levels of driving power, components operated well below their ratings, and the elimination of troublesome sliding contacts in tunable microwave cavity structures.

In addition, diode limiters are used to reduce changes in output amplitude level during normal tube aging. Routine maintenance will normally detect aged tubes before any resulting degradation is noted. All of the active circuits in the carrier supply are in tandem. A sudden failure of any one of these circuits along the frequency multiplier chains is detected at the 6301-mc output and initiates a fast switch. Additional detectors are located along the chains, as noted in the previous paragraphs, to assure that slow-level degradations will be detected and that appropriate switching action will result. These detectors control slow-acting Sensitrol relays.

Most of the components in the switching circuits themselves are re-

liable passive devices to insure low failure rates in the common output circuitry. The control circuits that operate the switches are also designed with long-life components which function at conservative operating limits. Manual switching controls permit routine maintenance without permanent loss of carrier. The switching circuits and their control circuits can also be maintained without loss of carrier.

7.2 *Low-Frequency Switching*

The low-frequency switching circuit is shown in much simplified form in Fig. 8. Terminations, crosstalk suppression paths, test points and monitor circuits are not shown. All leads are coaxial. The coaxial switches (223 type), controlled by the switching control circuit, are either all in the energized state when furnishing signals from the regular oscillator and low-frequency generator, or all in the de-energized state when furnishing signals from the standby equipment. The 29.7-mc and 59.3-mc outputs of the switching circuit are monitored to provide alarm information only. The transformer-switch arrangement shown allows a switch removal without totally disabling that carrier supply output. Test points are included on the transformers for in-service continuity testing.

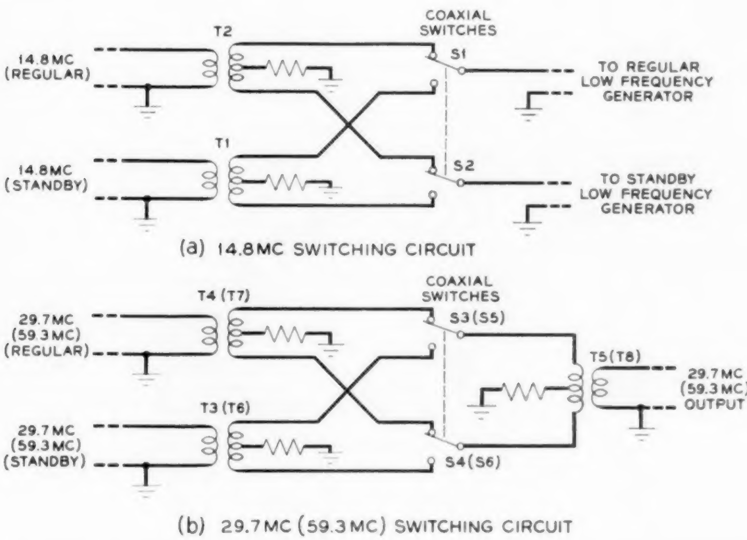

Fig. 8 — Simplified schematic of the low-frequency switching circuit.

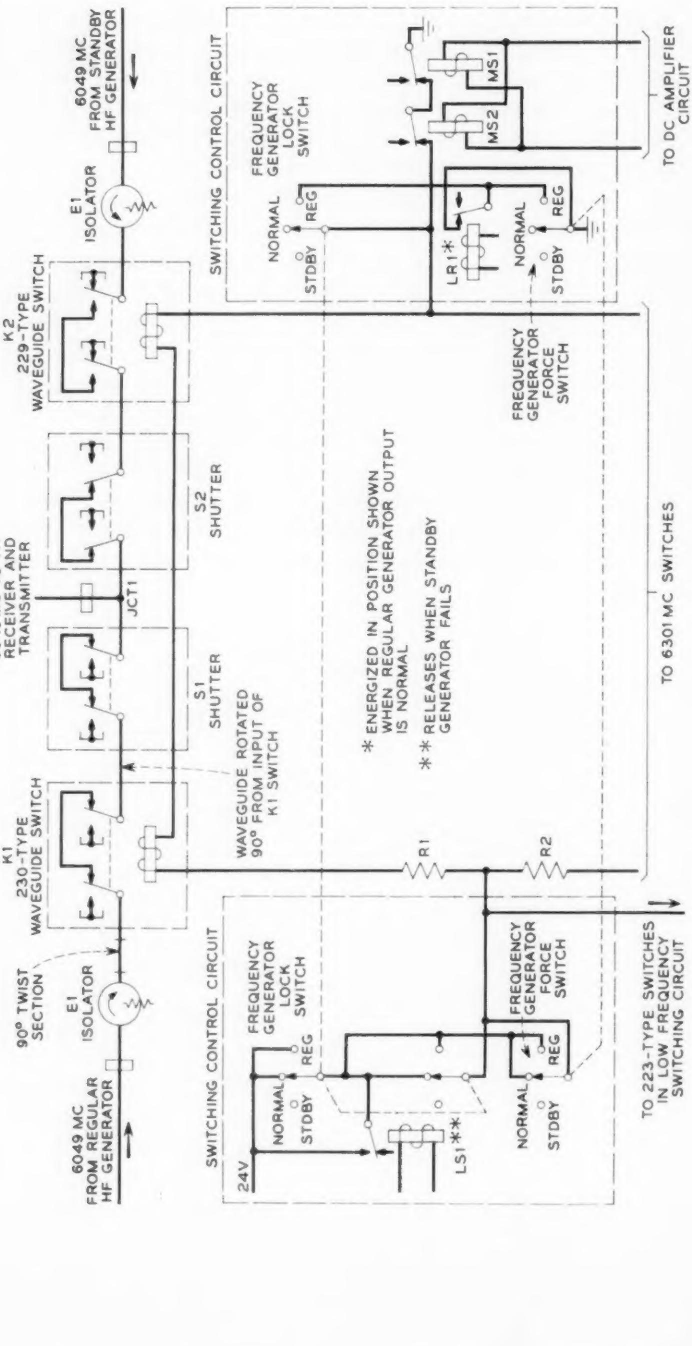

Fig. 9 — Simplified schematic for the high-frequency microwave switching circuit.

7.3 *Microwave Switching*

A simplified schematic of the microwave switching circuit is shown in Fig. 9. The microwave outputs of 6049 mc and 6301 mc are connected to the distribution circuit through ferrite-loaded microwave switches. Fig. 10 shows the polarization of the 6049-mc or the 6301-mc wave for the magnetized and unmagnetized conditions of one type of ferrite waveguide switch. The input and output rectangular waveguides are orientated to accept the polarization of the wave when the magnetic field is ON and to reflect the wave when the magnetic field is OFF. Another type of switch is identical except that the input and output waveguides are orientated to transmit the wave when the magnetic field is OFF and to reflect the wave when the magnetic field is ON.

The switches that transmit the signal when magnetized are connected to the 6049-mc and 6301-mc outputs of the regular high-frequency generator. The switches that reflect the signal when magnetized are connected to the outputs of the standby generator. The ferrite switch outputs are connected through a junction to the distribution network.

To provide fail-safe operation, the switch coils are connected in series as shown in Fig. 9 and are normally in the energized state. For this condition the regular generator supplies the distribution system, and the switches connected to the standby generator reflect the energy for

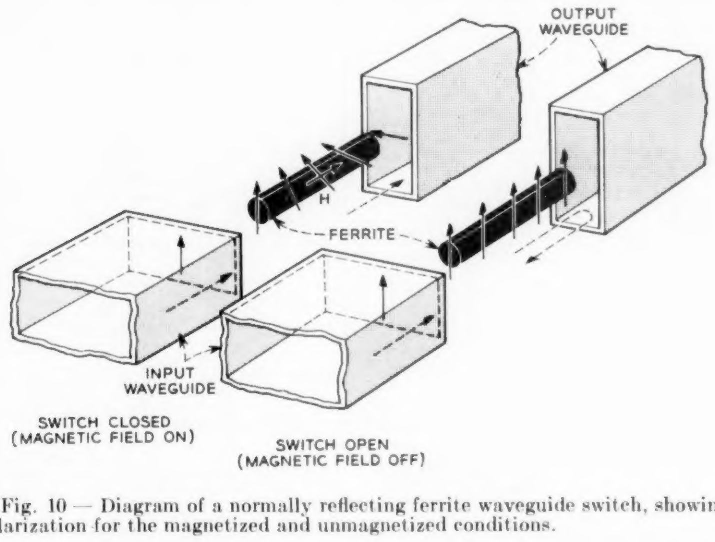

Fig. 10 — Diagram of a normally reflecting ferrite waveguide switch, showing polarization for the magnetized and unmagnetized conditions.

dissipation in isolators. For standby operation the coils of the switches are de-energized and the roles of the regular and standby switches are interchanged.

Manually operated shorting gates (shutters) are provided between the ferrite switches and the junction. Manual control switches (not shown on Fig. 9) which by-pass the coils of the normally transmitting ferrite switches are provided to facilitate removal of a ferrite switch under operating conditions. Automatic and manual control of the high-frequency switching circuit is performed by the switching control circuit.

7.4 *Switching Control*

A functional diagram of the switching control is shown in Fig. 11. Sudden equipment failures of the regular supply are detected at the 6301-mc output of the regular high-frequency generator. When this signal level drops approximately 4.5 db below the nominal value, the dc amplifier releases the 291-type buffer relays, which in turn release the coaxial relays in the low-frequency switching circuit and the microwave switches in the high-frequency switching circuit. The distribution circuit is now connected to the standby supply. Release of any one of the

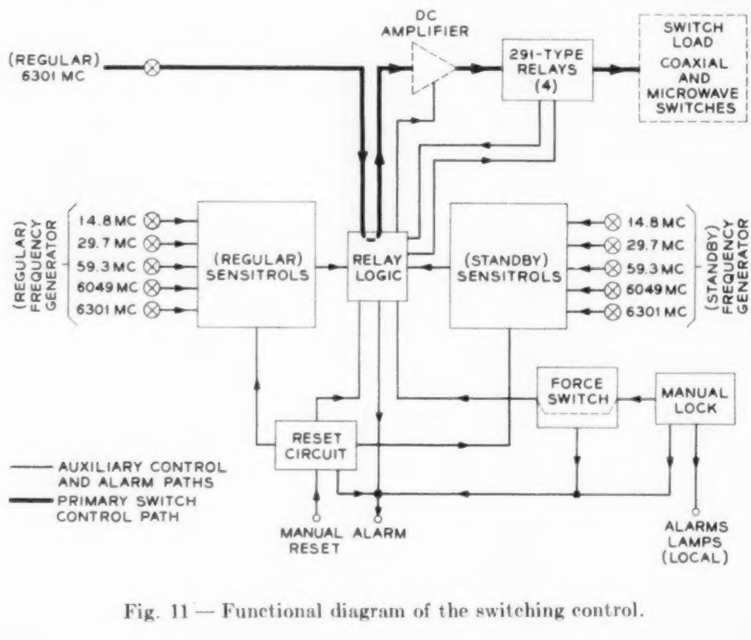

Fig. 11 — Functional diagram of the switching control.

Sensitrol relays connected to detectors at other points in the regular generator chain operates a relay which opens the dc amplifier input, causing a switch to the standby supply as before.

A magnetic amplifier, operating as a bistable switch, is the principal component of the dc amplifier. Its power source operates at approximately 20 kc to provide the required response time. The magnetic amplifier transfer characteristic is shown in Fig. 12. The power supply for the magnetic amplifier is a transistor inverter circuit consisting of a multiwinding transformer and two transistors. Failures of the magnetic amplifier, the inverter, or the power supplies associated with the dc amplifier result in a switch to the standby carrier supply.

The control circuits and switches are operated in a "fail-safe" manner; that is, failure in the control circuits or switches or loss of power to these devices will result in a switch to the standby supply.

The control circuit also energizes alarms and inhibits the operation of a manual switch to a failed supply. A forced switch is provided in the event a switch must be made to the "failed" supply. To prevent accidental forced switches, the forced switch operation involves the manual closure of two keys.

A lock-out circuit is provided to prevent the switching circuits from nigh-speed hunting when the switch is initiated by the 6301-mc monitor. The lock-out circuit short circuits the dc amplifier input once the dc amplifier has switched OFF, establishing a permanent switch to the

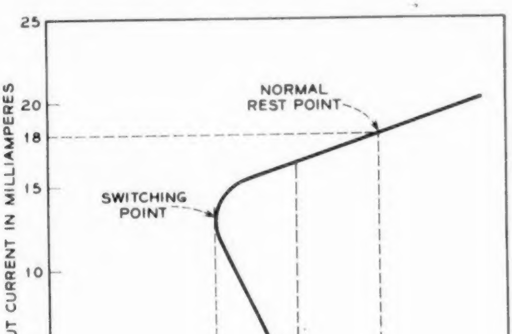

Fig. 12 — Typical transfer characteristic of the magnetic amplifier

standby generator. The lock-out is removed only after successful reset of the Sensitrol relays associated with the regular generator.

A reset operation is required to restore the regular supply connections to the distribution network. The reset may be accomplished manually or by remote control over the auxiliary channel facilities. The reset circuit consists of gas-tube timers and relays to reset the Sensitrol relays and to verify that the regular supply is operative before the switches are reset. The decision to permit a reset is based on information from the level detectors that connect to the regular Sensitrol relays, not by the ON switching point of the dc amplifier.

Detectors along the standby multiplier chain connect to the standby Sensitrol relays, which operate at level drops of approximately 4.5 db below nominal power output. Operation of a standby Sensitrol relay initiates alarms and prevents a manual lock to the standby generator. Manual locks can be performed only if the generator to which the lock is to be made is operative. However, once a successful lock is performed, the microwave carrier supply remains connected to that generator until the manual switch is cleared, regardless of the condition of that generator.

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FM Terminal Transmitter and Receiver For the TH Radio System

By E. W. HOUGHTON and R. W. HATCH

(Manuscript received June 12, 1961)

The FM terminals form an important subsystem of TH radio, as the link between the 0–10 mc baseband signal and the 74.1-mc FM signal. Severe requirements arise from the design objective of 16 terminal pairs in tandem in 4000 miles. The FM transmitter design uses a 6-kmc reflex klystron as a frequency modulator, the output of which is heterodyned down to 74 mc by another 6-kmc source. Automatic frequency control, with a 74.13-mc crystal oscillator as reference, provides the required frequency stability. In the FM receiver, an IF amplifier-limiter is followed by a balanced FM discriminator which uses parallel resonant discriminator networks. Improved linearity is obtained by special design of a common interstage network. The video amplifiers are balanced and use high-performance electron tubes. The over-all gain of a terminal pair is 8 db, between 124-ohm balanced video circuits.

I. INTRODUCTION

The relation of the FM terminals to the over-all TH system is described briefly in a previous paper.[1] There are two basic types of terminals: an FM transmitting terminal, which converts the baseband signal into a frequency modulated signal centered at 74.1 mc; and an FM receiving terminal, which recovers the baseband signal from the FM signal. The types of baseband signals to be transmitted[1] are shown in Fig. 1. In each terminal appropriate amplification is provided at both the intermediate and baseband frequencies to permit interconnection with other parts of the TH system. FM terminals are required at the ends of a TH route and at intermediate points where the baseband signal, or some portion of it, must be added or dropped. The design is based on a maximum of 16 terminal pairs in tandem in 4000 miles.

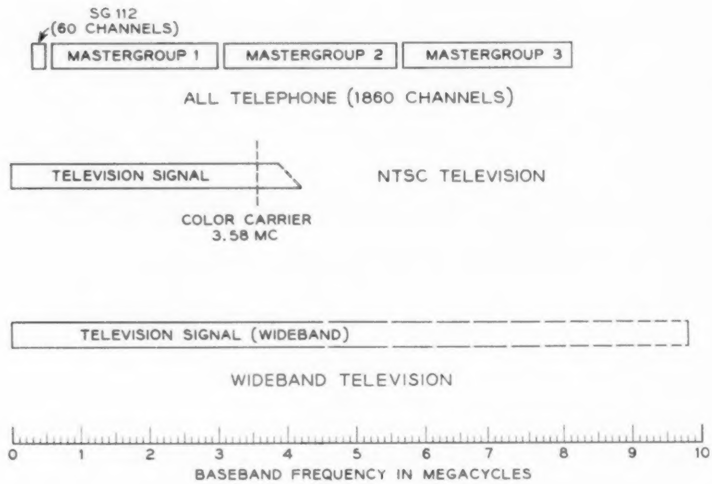

Fig. 1 — The three types of baseband signals transmitted in the TH system.

II. DESIGN OBJECTIVES AND CONSIDERATIONS

2.1 *Over-all Telephone and Television Objectives*

The over-all telephone (TP) and television (TV) objectives for 4000 miles of TH are based on extensive experience with previous systems. From these over-all objectives, allocations were made early in the TH development to the various portions of the system based on estimates of the relative difficulty and expense which would be required to achieve them. The allocations to the FM terminals are summarized in Table I.

Table I — Objectives for 16 Terminal Pairs in Tandem

Telephone (TP)	
Total noise (0 db TL)	31 dba
Fluctuation noise	28 dba
Cross modulation noise	28 dba
Television (TV)	
Weighted signal-to-noise ratio*	62 db
Differential phase	±1.25 degrees
Differential gain	±0.6 db

* The weighted signal-to-noise ratio is defined as the ratio of the peak-to-peak signal to the weighted rms noise, where the weighting is a function of noise frequency. For a detailed discussion see Ref. 2.

2.2 *Signal Characteristics*

Objectives such as those given in Table I can often be expressed in alternative ways. For example, the linearity objective necessary to restrict cross modulation noise is stated above in terms of the resulting noise in dba. For laboratory work, however, it is useful to state the linearity objective in terms of the harmonic performance when a sine wave of a given amplitude is applied. Conversions such as this depend upon certain system parameters. Assumptions, subject to some change when the over-all performance of the system could be determined, had to be made regarding signal characteristics such as the peak frequency deviation and the amount of pre-emphasis. The values of these quantities are shown in Table II.

2.3 *Objectives for an FM Terminal Pair*

Objectives for a single terminal pair were obtained by reducing the total allocation of Table I by a factor of either 4 (12 db) or 16 (24 db); the choice depends on whether a particular impairment could be expected to add on a random or on a systematic basis.

The principal design objectives for a single terminal pair are shown in Table III. In the sections which follow some of them will be discussed briefly.

2.4 *Baseband Transmission*

Essentially flat transmission is desired over the frequency band of all three types of signal. The upper frequency limit is approximately 10 mc for two; the lower frequency limit is set by the 60-cps component of a television signal. To provide adequate phase linearity for the low-frequency TV components, it is necessary to keep all low-frequency cutoffs considerably below 60 cps. Based on experience with other video systems, a low-frequency objective is expressed in terms of the distortion to a 60-cps square wave. A distortion not exceeding 2 per cent of the nominally flat top of the square wave is considered acceptable for a terminal

TABLE II — SIGNAL CHARACTERISTICS

Telephone signal	
Peak frequency deviation	4 mc
rms frequency deviation	0.7 mc
Pre-emphasis	7.5 db
Television only	
Peak frequency deviation	4 mc
Pre-emphasis (tentative)	12 db

TABLE III — DESIGN OBJECTIVES FOR ONE FM TERMINAL PAIR

Baseband transmission	
Bandwidth	2 cps to 10 mc
Gain stability	±0.25 db
Peak frequency deviation	4 mc
Center frequency of FM transmitter	74.13 ± 0.1 mc
Harmonic performance*	
Peak deviation for applied sine wave	4 mc
Second harmonic with respect to fundamental	−49 db
Third harmonic with respect to fundamental	−51 db
Differential gain*	See Note
Differential phase	0.3 degree
Fluctuation noise	
Message (at 0 db TL)	16 dba
Television (weighted signal-to-noise ratio)	74 db

* Differential gain and harmonic performance are analytically related as shown in Appendix A. The design objectives above for harmonic performance are set by cross modulation requirements for telephone; they also insure adequate differential gain performance for television.

pair. To meet this, the low-frequency cutoff (3-db point) actually occurs at about 2 cps. Transmission flatness of about ±0.1 db is the objective for the band from 60 cps to 10 mc.

The objective of ±0.25 db for gain stability comes from two main considerations. First, there is the desire to control net loss in toll telephone channels within rather close limits. Second, all operating terminals at a given point must have almost the same net loss as standby terminals. Otherwise, switching from a regular terminal to a protection terminal will cause hits in data signals.

2.5 *Harmonic Performance*

Nonlinearity in the terminals causes cross modulation in the TP signal, and differential phase and gain in the TV signal. It was found that the linearity required for the telephone signal was controlling by a slight margin. This led to the objective on harmonic performance given in Table III.

2.6 *Differential Gain and Phase*

The NTSC color television signal uses a modulated 3.58 mc carrier to transmit color information. Intermodulation between the low-frequency luminance information in the signal and the color carrier causes amplitude and phase variations in color carrier which are a function of the luminance signal. These variations show up as distortion in the saturation and chroma of the reproduced TV picture.

A special test signal[3] consisting of a low-level 3.58-mc tone and a higher-level 15.75-kc tone is used to simulate the TV signal and to test for distortion of this type. Variations in the amplitude and phase of the 3.58-mc tone as a function of the 15.75-kc tone are referred to as differential gain and phase. For tests of this type the peak-to-peak amplitude of the low-frequency tone is normally made equal to the peak-to-peak amplitude of the TV signal it simulates. It is then possible to specify quantitatively the amount of differential gain and phase which corresponds to a tolerable amount of color distortion in the TV picture.

Since differential gain and phase distortion occurs as the result of intermodulation, the amount of distortion which occurs in a particular nonlinear system depends on the amplitude of the applied signal. Differential phase and gain distortion can be reduced by reducing the signal amplitude, but at the expense of a poorer signal-to-noise ratio. A compromise between these two types of signal degradation is often possible by reducing the amplitude of the low-frequency components of the TV signal before transmission by means of a pre-emphasis network. An inverse network is used to compensate at the receiving end. Pre-emphasis of the amount shown in Table II is proposed for this system. With this pre-emphasis, the objectives for harmonic performance shown in Table III ensure that the over-all differential gain objectives will be met.

III. FM TRANSMITTER

The FM transmitter provides a +11-dbm output signal, centered at 74.1 mc, which is frequency modulated proportional to the input baseband signal. One volt peak-to-peak signal produces 8-mc peak-to-peak frequency deviation.

To preclude excessive intermodulation, very little nonlinearity is permitted in the frequency deviation vs voltage characteristic. This objective had a very strong influence on the selection of a modulation method. Other objectives that had important influences on detailed design approaches were: the wide baseband, a conversion gain stability of ±0.15 db, and a carrier frequency (74.13 mc) stability of ±0.1 mc.

After a considerable exploratory development period during which several more compact circuits were rejected because of marginal linearity or bandwidth, a reflex klystron was selected as the frequency modulator. A similar circuit using klystrons is successfully employed in the FM modulator associated with the TD-2 radio system.[4] However, improvements in the klystron and its associated circuits were essential to meet the more stringent requirements of the TH system.

3.1 *General Description*

Selection of a klystron modulator essentially establishes the principal accessory units. These are shown on Fig. 2.

The baseband signal, amplified by the video amplifier and applied to the repeller of the deviated (DO) klystron, causes its frequency to vary around a rest value of approximately 6174 mc. The deviated signal is applied through an isolator to the converter, where it is mixed with a 6100-mc signal from the beating (BO) klystron. The output, a frequency modulated wave centered on 74.1 mc, passes through a delay equalizer (which compensates for the delay distortion of a tandemly connected FM transmitter and FM receiver), to an IF amplifier. The amplifier output connects via coaxial cable to the other parts of the TH system.

To provide an input for the automatic frequency control (AFC) circuit, a small fraction of the output from the FM transmitter is abstracted and amplified. Alternating samples of this FM signal and the output of a crystal-controlled 74.1-mc oscillator are applied to the limiter-discriminator. The output is a square wave with an amplitude proportional to the frequency difference. This error signal is amplified and rectified in the synchronous detector. The resultant voltage is applied to the BO klystron with the proper polarity to reduce the average frequency error.

A photograph of the FM transmitter is shown in Fig. 3. The klystrons, converter and other microwave devices are mounted in the FM generator panel. The video amplifier is at the bottom and the transmitting IF amplifier is just above the frequency comparator panel. A further description of the equipment features is given in a companion paper.[5]

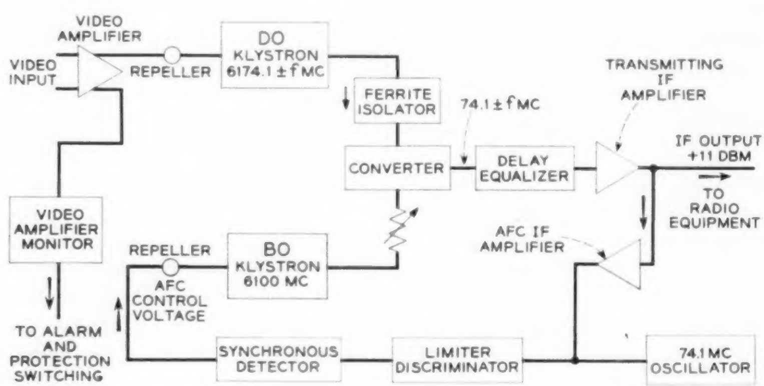

Fig. 2 — Block diagram of FM transmitter.

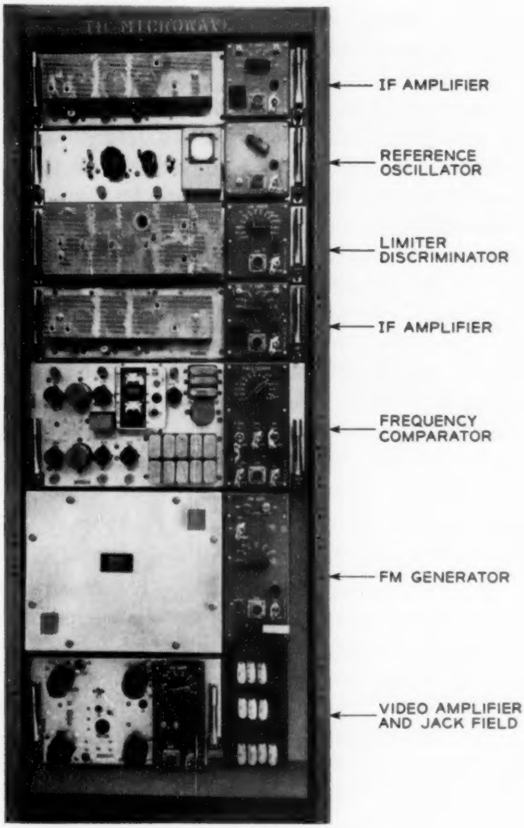

Fig. 3 — The FM terminal transmitter.

3.2 *Video Amplifier**

The gain of the video amplifier is such that a signal of 1 volt peak-to-peak applied to its 124-ohm balanced input is increased to 5.5 volts peak-to-peak at the repeller electrode of the DO klystron, which is sufficient to deviate the klystron 8 mc peak-to-peak. At this output voltage, a typical video amplifier has second and third harmonics that are below the fundamental by 60 db and 75 db respectively.

To realize this performance, two balanced electron-tube amplifier

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* Important contributions were made to this and other sections dealing with the video amplifiers by H. C. Hey.

stages are used in a circuit configuration similar to that of the video amplifier described in Section IV. Low second- and third-harmonic distortion results from the use of relatively lightly driven, high-current electron tubes in the two stages. Western Electric 448A tetrodes and 437A triodes are used, respectively, in the input and output stages. Additional suppression of the second harmonic is achieved through the balancing action of push-pull stages, which have cathode feedback to reduce and stabilize residual unbalances.

To keep the gain-frequency distortion less than ±0.1 db up to 10 mc, design emphasis is placed on minimizing spurious interstage capacitance. Interstage resistances are limited to values which give a 3-db gain reduction at 14 mc. Interstage compensation, part of which is individually adjusted, is used to achieve flat gain. The use of large coupling capacitors, along with some additional phase compensation, limits 60-cps square wave distortion to less than 1 per cent.

3.3 *Microwave Circuit*

The microwave circuit consists of the two klystrons, an isolator, an RF attenuator and a converter. The converter employs a silicon diode in an unbalanced microwave network with internal resistance padding to reduce changes in performance with different diode characteristics. The output power from the klystron is sufficiently high to make unimportant the added conversion loss due to the padding. Variations in the output amplitude of the FM wave are reduced by the limiting action obtained in the converter. This limiting action is achieved by making the power from the frequency-deviated oscillator (DO) substantially higher than that from the undeviated beating oscillator (BO). The RF attenuator determines the level difference, since the two klystrons are of the same type and generate approximately the same power output. Presentation of a well-matched impedance to the DO klystron is required for modulation linearity, and is obtained by use of the microwave ferrite isolator, as shown in Fig. 2.

Radiated microwave interference to or from the klystrons is precluded by enclosing them in a shielded compartment into which all leads are brought through microwave filters.

3.4 *Reflex Klystron Modulator*

To meet the stringent linearity objectives for the TH system a new klystron, coded as the Western Electric 450A, was designed. It has a limited tuning range (6000 mc to 6200 mc), a low-Q resonator (loaded

$Q \approx 100$) and requires resonator, repeller and heater potentials of approximately 450, -100 and 6.3 volts, respectively.

The loaded Q of the klystron is important in determining both the linearity and the fluctuation noise performance. As shown in Appendix B, the principal distortion terms due to nonlinearity in the voltage-frequency deviation characteristic are proportional to Q^2. On the other hand, fluctuation noise, due to shot noise in the electron stream, is proportional to $Q^{-1/2}$ (Appendix D). Thus, the selection of too high a Q leads to excessive intermodulation products, whereas too low a Q causes excessive fluctuation noise. A compromise value of 100 was selected to give the best over-all performance. With this, the second and third harmonics are respectively at least 54 db and 56 db below a fundamental which has a peak-to-peak deviation of 8 mc. The FM components of the fluctuation noise, as measured by an FM receiver, are flat with frequency above about 25 kc. Below 25 kc the noise power increases with an approximate $1/f$ law. Above 25 kc the rms frequency deviation due to the noise in two klystrons is approximately 0.6 cps in a one-cycle band.

The deviation sensitivity of the klystron is approximately inversely proportional to loaded Q, and essentially independent of baseband frequency up to 10 mc. The reasons for this are demonstrated in Appendix C.

3.5 *IF Amplifier*

The two identical IF amplifiers (transmitting and AFC) shown in Fig. 2 each have three tubes. Each has a minimum gain of 21 db, a maximum output of $+11$ dbm, and a bandwidth of 58 to 90 mc between 0.3-db points. The bias on the intermediate tube, supplied from an external source, can be varied for manual gain adjustment or for gating the amplifier on and off as required in the AFC circuit.

Input, output and intermediate stages in this amplifier are almost identical in electrical design to corresponding stages in the main IF amplifier of the radio receiver, described in a companion paper.[6]

3.6 *AFC Circuit*

The AFC circuit, shown schematically on Fig. 4, is designed to hold the average frequency of the outgoing FM wave at 74.1 ± 0.1 mc. A low-temperature-coefficient, crystal-controlled oscillator operating at 74.130 mc provides the basic reference against which the average frequency of the outgoing FM wave is compared. By alternately gating the 74.1-mc reference oscillator and the AFC IF amplifier on and off at a

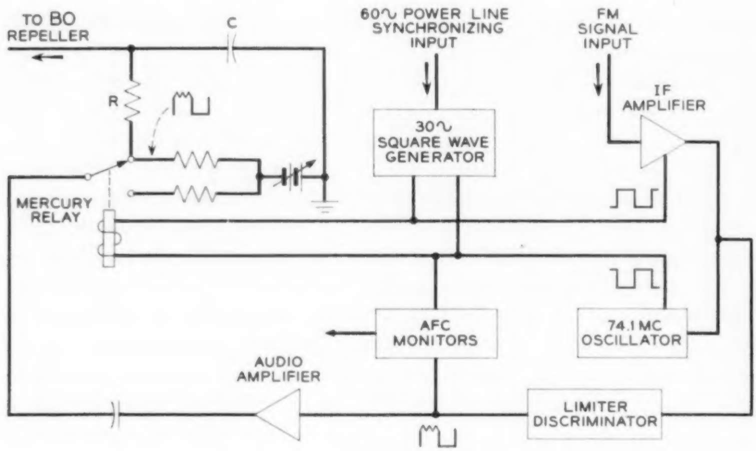

Fig. 4 — Functional schematic of the AFC system.

30-cps rate, there appears at the IF limiter input during one $\frac{1}{60}$-second interval a signal from the 74.1-mc oscillator, and during the next $\frac{1}{60}$-second interval a sample of the outgoing FM signal. The limiter section equalizes the signal amplitudes, and the frequency discriminator delivers to the audio amplifier a 30-cps square wave whose peak-to-peak average amplitude is proportional to the frequency difference between the reference oscillator and the average (carrier) frequency of the outgoing FM wave. After amplification the 30-cps wave is rectified in a synchronous detector (the mercury relay), the dc output of which is related in amplitude and polarity to the frequency difference. After filtering, this dc is applied to the repeller of the BO klystron with a polarity which causes its frequency to change in a direction to reduce to a small value the difference (error) in frequencies of the 74.1 mc oscillator and the outgoing FM wave.

Since the instantaneous frequency of the FM wave is continuously changing, nonlinearity in the AFC discriminator will give an average output not linearly related to the average frequency of the FM wave. This discriminator has a first-order nonlinearity of 2 per cent or less between 70 and 78 mc. On the basis that the rms frequency deviation will be less than ± 1 mc, the average frequency shift error will be less than 5 kc.

The use of ac amplification following the discriminator makes the circuit insensitive to small drifts in the center frequency (corresponding to

a dc output of zero) of the discriminator. Synchronous rectification of the amplified ac output restores sense so that the resultant voltage is applied to the BO repeller with the correct polarity. Since the AFC open-loop gain is typically 40 db, uncorrected frequency differences of as much as 10 mc are reduced by AFC action to 100 kc.

When the outgoing FM wave contains a television signal, large 60-cps and harmonically related voltages are present in the discriminator output during signal sampling intervals. A dc output related to the true average values of the signal voltage must be obtained for accurate frequency control. This is accomplished by using negative feedback to linearize the audio amplifier and by using a mercury relay switch for rectification. The switch has negligible storage reactance in its output load so that its dc output is proportional to the average value of the input.

When the frequency of the local TH ac power differs significantly from that of the remote power line against which the television signal is synchronized, successive signal samples begin (and end) at different phases of the 60-cycle ac component in the TV signal. Consequent "flicker" interference[7] on television signals caused by the apparent shifts in average frequency at the beat frequency rate are reduced to tolerable values by the RC filter following the rectifier.

Other beat frequency effects are kept small by synchronizing the 30-cps gating and rectification functions with local power frequencies. The RC filter on the rectifier output has a 3-db cutoff at 0.005 cps. It gives 60-db loss at 5 cps, the most annoying beat frequency, and reduces the 40-db AFC loop gain to unity at 0.5 cps. Transient response is optimized by virtue of the 90° phase asymptote. Video phase shift, and consequently the 60-cps square wave response of the FM transmitter, is only slightly affected by the AFC action, which has an effect approximately equal to that of a low-frequency cutoff at 0.5 cps.

IV. FM RECEIVER

The FM receiver accepts the 74-mc FM signal and delivers a balanced baseband output signal which is 8 db above 1 volt peak-to-peak in a 124-ohm circuit for 8-mc peak-to-peak frequency deviation.

Performance requirements on linearity, baseband transmission, stability and noise are comparable to those already discussed for the FM transmitter. After careful study, the design described in the following sections was selected as the best compromise between over-all performance and ease of maintenance and adjustment. In its main features the design is similar to that which has been used in the TD-2 radio system[4] for a number of years. Modifications in the detailed circuitry, however, have

achieved a considerable improvement in bandwidth, linearity and circuit stability.

4.1 *General Description*

A block diagram of the FM receiver is shown in Fig. 5. The FM input is applied to the receiver by means of 75-ohm coaxial cable. Provision is made for an IF amplifier, identical with the two in the FM transmitter, in case additional IF gain is required. In any event, the FM signal at +1 dbm is applied to an amplifier-limiter to suppress any amplitude modulation of the signal, which otherwise would cause unwanted distortion in the discriminator. In addition, the limiter action tends to maintain a constant input power for the discriminator circuit. Without it, the discriminator output would vary linearly with changes in input carrier power to the FM receiver. This would cause undesirable variations in the net gain of a terminal pair. The amplifier-limiter is electrically identical to the one in the broadband radio transmitter.[6] The mechanical design, however, is somewhat different,[5] to be in keeping with the plug-in styling of the other FM terminal circuits.

The adjustable attenuator ahead of the discriminator is used to set the sensitivity of the discriminator section to a standard value. The discriminator recovers the baseband signal from the FM wave, in two steps. First, networks which introduce amplitude slope across the IF band produce amplitude modulation, which is proportional to the frequency modulation of the input signal. Amplitude detectors then recover the baseband signal, which is subsequently amplified in the baseband amplifier.

A photograph of an FM receiver is shown in Fig. 6, in which the individual units are easily identified.

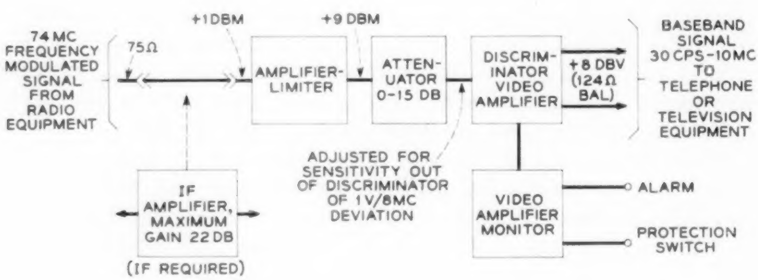

Fig. 5 — Block diagram of FM receiver.

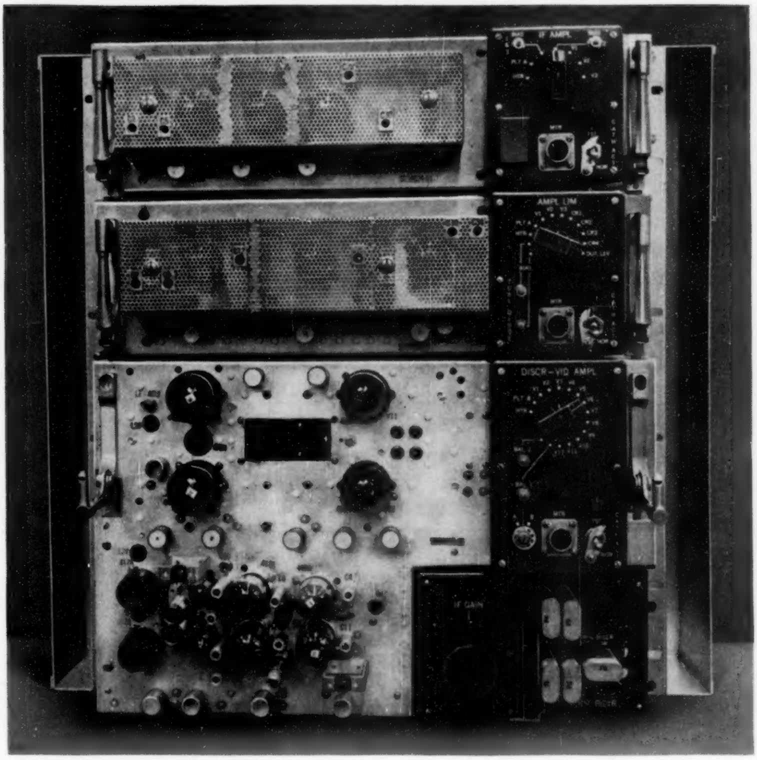

Fig. 6 — The FM terminal receiver.

4.2 *FM Discriminator and Video Amplifier*

Simplified schematics of the FM discriminator and of the video amplifier are shown in Figs. 7 and 8. The input signal is transformer-coupled to two 448A electron tubes, V_1 and V_2, which are operated in parallel. Their combined output is developed across a common interstage, consisting essentially of a parallel resonant circuit, and applied to the grids of the 418A electron tubes, V_3 and V_4, which are driven in parallel. All four tubes are stabilized by dc feedback and cathode compensation networks (Z_1—Z_4) as described for the IF amplifier in the broadband radio receiver.[6] The common interstage performs an important function in the over-all design which will be described later. A 75-ohm test jack is provided to aid in interstage adjustment.

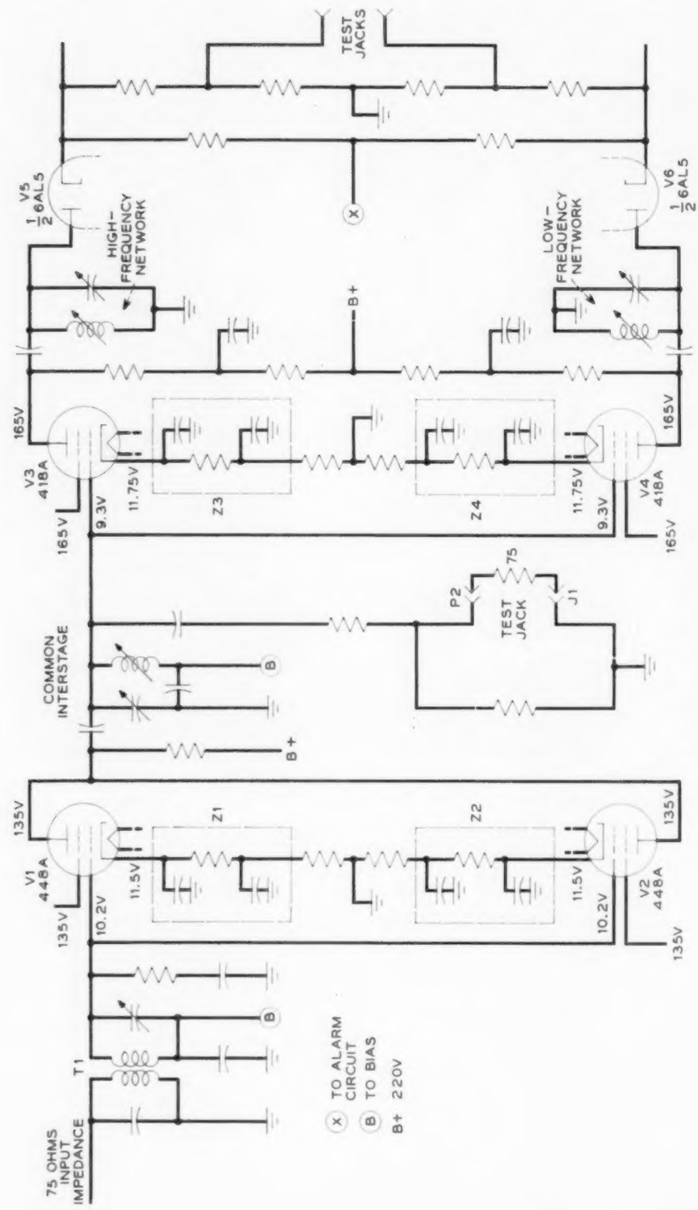

Fig. 7 — Simplified schematic of FM discriminator.

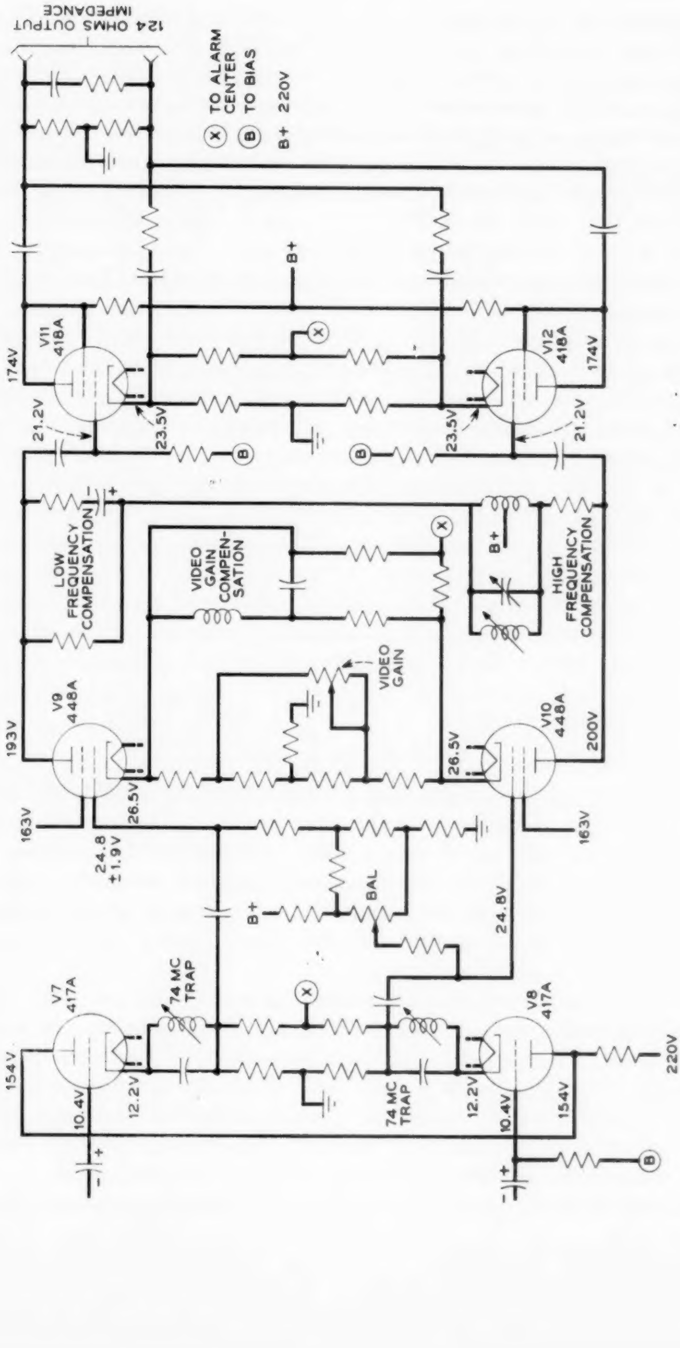

Fig. 8 — Simplified schematic of video amplifier.

Tubes V3 and V4 are used as constant current generators to drive separate parallel-resonant discriminator networks. The high-frequency network is peaked at approximately 100 mc, and the low-frequency network is peaked at approximately 50 mc. Each provides a large amplitude slope across a wide band centered at 74.13 mc. Tuned at these frequencies, the slope of one is positive and the other negative across the IF band. The output voltages developed across the networks are amplitude modulated with 180° phase difference between them. The diode detectors, V5 and V6, recover the baseband signal from the amplitude modulation, thus providing a balanced input to the balanced video amplifier which follows.

The video amplifier section (Fig. 8) consists of three balanced stages in tandem. The first is a cathode follower stage employing 417A triodes. The cathode follower is used to minimize the capacitance facing the diode detectors. Parallel resonant traps are provided in the cathode circuits to reduce the amount of 74-mc carrier entering the amplifier stages which follow. The second stage, with 448A tetrodes, provides the required voltage amplification and feeds the balanced output stage. Fixed low-frequency compensation and adjustable high-frequency compensation are provided in the interstage. The output stage, using 418A tetrodes, is connected as a modified cathode follower circuit. The output is increased by connecting both cathodes and plates to the load in an arrangement which is used in the A2A television transmission system.[8]

4.3 *Design Considerations*

To meet the over-all system objectives, the discriminator has to be very linear, yet give an output large enough to provide an adequate signal-to-noise ratio at the input to the video amplifier. Furthermore, the circuit must be readily adjustable to allow for manufacturing variations and subsequently be stable with time. The manner in which the design has been affected by these considerations is discussed in the following paragraphs, in which the usual order is reversed by working from the output toward the input.

A study of the noise and microphonics expected in the first stages of the video amplifier leads to the establishment of an objective of 1 volt peak-to-peak at the detector output, for a peak-to-peak frequency deviation of 8 mc. This level is sufficient to keep the fluctuation noise contribution of the FM receiver at least 10 db below that of the FM transmitter and at the same time to prevent microphonics in the first stages of the video amplifier from degrading the television signal.

The use of 6AL5 diodes for the AM detectors provides a compromise

among the following objectives: good linearity, high detection efficiency and ease of replacement. The operation is between that of an averaging detector and a peak detector. The over-all detection efficiency is approximately 40 per cent, close to that of an ideal averaging detector. Thus, the capacitance in the output circuit for the diodes provides some peaking action to compensate for the loss due to the forward resistance of the diode. However, this capacitance must be kept low to minimize video roll-off at 10 mc.

From the desired output of 1 volt peak-to-peak and the diode efficiency of 40 per cent, the necessary change in IF signal amplitude as it is tuned across an 8-mc band is about 1.25 volts for each discriminator network. This change in signal amplitude is a function of the discriminator networks and the signal currents provided by the preceding tubes. Restrictions are imposed by the linearity objective and the interstage capacitance which must be absorbed. This limits the maximum change in impedance which can be achieved across the 8-mc band to about 50 ohms. Thus, peak signal currents of about 25 ma are required from the driving tubes. Furthermore, this amplitude must be provided with low harmonic content. The second harmonic, in particular, will be enhanced with respect to the fundamental by the amplitude-frequency characteristic of the high-frequency network. The detector output will therefore contain an error term due to the harmonics. Good harmonic performance is required to permit accurate adjustment of the discriminator with sweep signals, and to a somewhat lesser extent, to prevent distortion to the normal signal.

The need for a large signal current with low harmonic content led to the selection of the 418A tube for this application.

4.4 *Discriminator Linearity*

The discriminator networks have a substantial amount of curvature, predominantly parabolic, as shown in Fig. 9. This curvature, if uncompensated, would result in a nonlinear relationship between the incoming frequency modulation and the resulting amplitude modulation. One method of correction is to use more complex discriminator networks. This was not selected because of the difficulty in controlling parasitic capacitance and inductance. A second approach, used in the discriminator for the TD-2 system, is to select network designs such that the parabolic curvatures of the two sides are equal. The predominant second-order modulation products then tend to cancel each other in the balanced output from the detectors. The major difficulty with this approach is the amount of balance required. An analysis of the networks in the

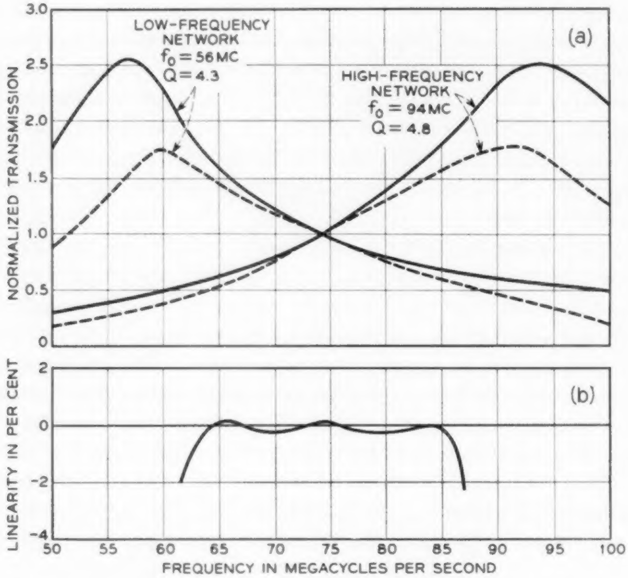

Fig. 9 — (a) Discriminator network characteristics; dashed curves show effect of common interstage; (b) typical discriminator linearity.

TH discriminator indicated that the second-order modulation products from one side of the discriminator would be about 30 db above the desired objective. Although a 30-db balance is obtainable at the time of adjustment, it is difficult to hold between maintenance intervals since it depends on the relative gains of v3 and v4.

Another approach,* the one used in this design, also requires the networks to have equal parabolic curvatures. However, instead of depending on precise cancellation in the output, a compensating parabolic shape of opposite sign is introduced in the common interstage ahead of V3 and V4. In this way the linearity of each side of the discriminator is substantially improved, as shown in Fig. 9. With this design the nonlinearity of each side exceeds the over-all objective by only about 10 db. This reduces by a considerable amount the effect of gain changes in the driving stages v3 and v4. For example, without the common interstage, only 0.5 db of change in the relative gains after the initial balance would

* This approach was suggested by N. E. Chasek of the Radio Research Department at Bell Telephone Laboratories.

cause the objective to be exceeded. With the common interstage, the acceptable variation is 4 db. Thus, the common interstage substantially increases the time stability of the second harmonic balance.

The procedure used in designing the discriminator networks is outlined in Appendix E. This design was modified slightly on the basis of experimental results to obtain the typical linearity characteristic shown in Fig. 9(b).

V. POWER SUPPLIES

Supplied with regulated, reliable 220-volt, 60-cycle ac inputs, the power rectifiers give dc voltages of -11, $+135$ and $+220$ volts for the FM receiver; -11, -170, $+135$, $+220$ and $+450$ volts for the FM transmitter. Electronic regulation is counted on to reduce bobble and consequent television flicker interference to acceptable levels. This stability also keeps within limits, for extended time intervals, variations in FM deviation, FM sensitivity, and second harmonic balances. Additional details of the power supplies are given in Ref. 9.

VI. MONITORS

To preclude disruption or excessive degradation in service due to failure in an FM transmiter or an FM receiver, the failed unit is automatically replaced by a standby unit. Status information for initiating this protection switching action and for registering alarms is obtained from monitors on space current in video amplifier tubes, on IF carrier power at the FM transmitter output, on three significant parameters in the AFC system, and on rectified carrier level in the FM receiver.

6.1 *Video Amplifier Monitors*

A considerable simplification in instrumentation of video amplifier monitors is based on the fact that any reduction in gain, or increase in harmonic distortion, will probably be accompanied by a change in space current in one or more electron tubes. Cathode voltages (which are proportional to space currents) from all tubes are added, and the sum is applied to the input of a differential dc amplifier. A change in this sum resulting from a change in any cathode voltage of 30 per cent or greater will initiate an alarm and a protection switching order. Even though the response time of this monitor is around two milliseconds, it is not operated by television waveforms having frequency components as low as 60 cps because of the push-pull action in the balanced video amplifier.

6.2 *IF Carrier Monitors*

Failures in klystrons or the transmitting IF amplifier in the FM transmitter are detected by a reduction in output from an IF carrier level detector connected across the outgoing IF line.

By monitoring the sum of rectified voltages at the discriminator output in the FM receiver, information is obtained on the status of IF carrier input and of the electron tubes and circuits in the discriminator. This voltage sum is combined with the sum of the FM receiver video amplifier cathode voltages and applied to a video amplifier monitor circuit of identical design to that used in the FM transmitter.

6.3 *AFC Monitors*

Because of the large time constant in the output filter, a failure in the AFC control circuit will not cause an immediate change in the average frequency of the outgoing FM wave. This allows the use of three reliable, though relatively slow-acting, sensitive meter relays for monitoring this circuit. These relays monitor: (a) the peak-to-peak frequency error voltage, (b) the rectified carrier level, and (c) the 30-cps gating voltage. Failures in the audio amplifier or synchronous rectifier will cause monitor (a) to initiate alarm and protection switching orders whenever the klystrons drift sufficiently to create a 1.5-mc shift in difference frequency. However, monitor (a) is a null indication and so will not be activated by failures in the gating circuits or in the limiter-discriminator and preceding IF circuits. Failures in these circuits are detected by monitors (b) and (c).

VII. TANDEM PERFORMANCE

The baseband-to-baseband performance of a typical FM transmitter and FM receiver when connected in tandem is discussed in this section. This performance, compared with values in Table III, will show how well the design objectives have been met.

It should be noted that "typical" back-to-back performance means the average performance that can be expected during the time intervals between maintenance adjustments on individual FM transmitters and receivers. These specifications draw from experimental data in attempting to give the most probable performance that can be expected over a reasonable time interval after maintenace adjustments have been made. They observe the practical restriction that terminals cannot be connected back-to-back in the field to adjust paired terminals for optimum performance.

7.1 *Envelope Delay Distortion (EDD)*

Measured as the change in phase shift of a 278-kc tone while the IF center frequency is swept between 64 mc and 84 mc, the IF delay distortion (excluding the delay equalizer shown in Fig. 2) has the typical characteristic shown on Fig. 10(a). Odd-order EDD at 64 and 84 mc is approximately -5.3 and $+5.3$ mμs respectively, and even-order EDD is approximately $+8.7$ mμs at both frequencies. To this total the klystron and its microwave circuits contribute almost nothing, the FM discriminator contributes about half, and the remaining IF circuits (amplifier-limiter and transmitting IF amplifier) contribute about half. With the delay equalizer, the distortion is reduced by a factor of at least five.

7.2 *Harmonic Distortion*

Typical curves of second and third harmonic performance (exclusive of the delay equalizer) are shown in Fig. 10(b). At low frequencies the harmonic performance is flat with frequency, and is primarily due to nonlinear voltage-frequency characteristics in the klystron and in the discriminator. At higher frequencies, however, second and third harmonics tend to increase in proportion to the frequencies where the harmonics fall. This is primarily due to EDD in the IF circuits. This source is reduced by delay equalization until it is negligible compared to the low-frequency asymptote. The performance then becomes that shown by the dash lines in Fig. 10(b).

The limiting low-frequency value of -70 db for the third harmonic-to-fundamental ratio comes from systematic contributions by two video amplifiers, the DO klystron and the FM discriminator; typical values are respectively -0.0004 (-68 db), $+0.0015$ (-56 db) and -0.001 (-60 db). The second harmonic contribution from the same four units depends upon the status of balance in each unit. The low-frequency asymptote of -50 db was obtained by taking the root-sum-squares (R.S.S.) of limiting design values of ± 0.001 (-60 db), ± 0.002 (-54 db) and ± 0.002 (-54 db) in each of two video amplifiers, the DO klystron and the FM discriminator, respectively.

7.3 *Differential Gain and Phase*

As discussed previously, the harmonic performance given above insures adequate differential gain performance for television.

Without delay equalization, the differential phase characteristic has the same broad structure shape as the EDD characteristic of Fig. 10(a),

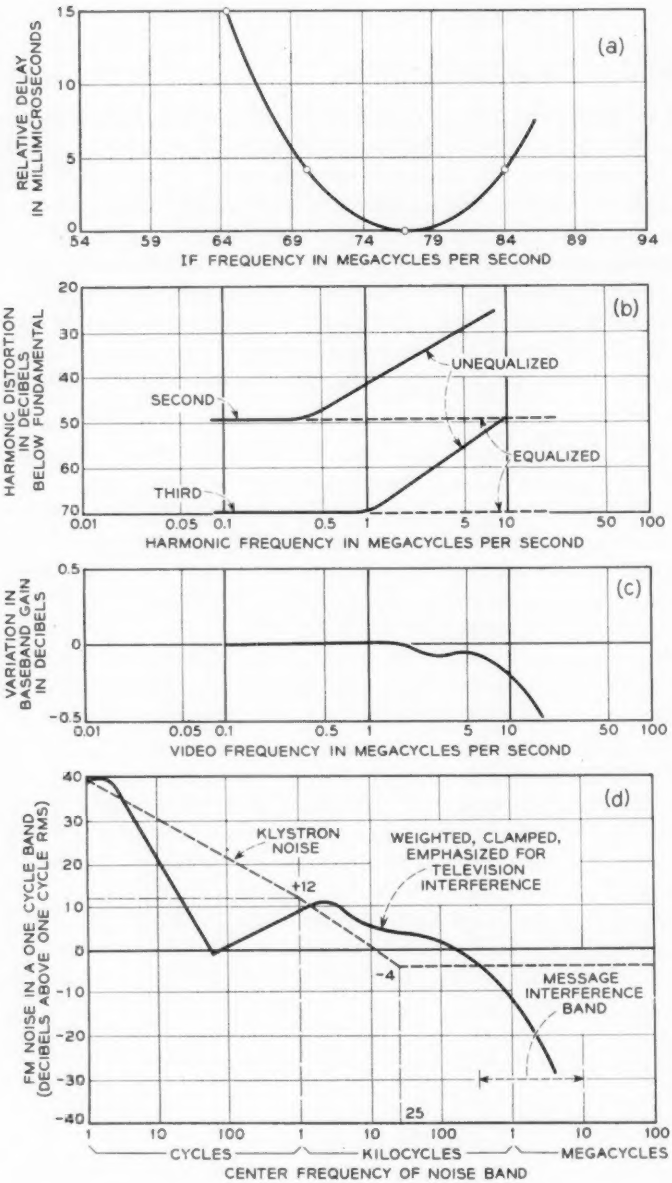

Fig. 10 — Tandem characteristics for an unequalized pair: (a) IF delay distortion, (b) typical harmonic distortion for 8-mc peak-to-peak deviation, (c) typical baseband response, (d) fluctuation noise.

but with reversed sign, since it is also the change in phase shift (in this case at 3.58 mc) while the carrier is swept by a low-frequency (15-kc) video voltage. Odd-order and even-order differential phase are each about 2.5° at ±4-mc peak deviation. Since the low-frequency components of the TV signal are less than this by 12 db due to pre-emphasis, the odd- and even-order differential phase experienced by the TV signal is reduced by factors of at least 4 and 16 respectively. Differential phase is further reduced by the delay equalization, and hence is well below the 0.3° objective.

7.4 *Baseband Amplitude Response*

Low-frequency baseband response, determined by the characteristics of the two video amplifiers, is typically down 3 db at 2 cps. This, with some added low-frequency phase equalization, gives a 60-cps square wave response having low-frequency slope of less than 2 per cent. High-frequency baseband response, typically ±0.15 db at 10 mc, is the result of a number of significant contributions.

As described in previous sections, video response in the klystron and its microwave circuits is substantially constant. Response in each video amplifier is adjusted to be constant up to 10 mc within ±0.1 db.

The EDD of Fig. 10(a) produces a video roll-off of 0.06 db at 10 mc. In contrast to harmonic distortion, amplitude response is made worse by delay equalization when the distortion and the equalization are separated by a limiter. Thus, equalization in the FM transmitter, which introduces an equal and opposite delay distortion, adds another 0.06 db to the video roll-off at 10 mc.

Contributing directly to baseband gain-frequency variations, even-order IF gain variations in a typical transmitting IF amplifier account for ±0.03 db. Though somewhat more subtle to observe experimentally because of limiting action, the same basic mechanism in the amplifier-limiter contributes approximately the same video gain deviation. Linearity and detector respo[illegible] in the FM receiver results in a systematic even-order characteristic [illegible]bout −3 per cent at 64 mc and 84 mc, contributing −0.3 db to the video response at 10 mc.

Summing up the systematic contributions and adding the random contributions on an rss basis leads to a typical characteristic of −0.4 ± 0.15 db at 10 mc. The systematic characteristic is equalized to give the nearly flat response shown on Fig. 10(c).

7.5 *Low-Frequency Noise*

Noise below 300 kc causing interference in TV signals comes from mechanical vibration in electron tubes, from "$1/f$" cathode emission

fluctuation noise in klystrons, from power rectifier output voltages having "hum" components which are multiples of power-line frequency, and from "bobble" components which are random variations in power rectifier dc outputs at rates of 1 to 30 cps.

Rugged mechanical design in the klystrons and maintenance of relatively high signal levels at the FM receiver video amplifier input minimize mechanical vibration noise. Considering frequency weighting, TV circuit clamper characteristics, and 12-db pre-emphasis for the combined service, the contribution by "$1/f$" cathode emission noise over the band of 0 to 25 kc can be neglected since it is more than 6 db below weighted rms fluctuation noise in the rest of the TV band.

From the standpoint of deriving objectives and evaluating performance, an especially troublesome problem centered around the inevitable small, random variations in the 60-cps ac power. This power line "bobble" creates "flicker" interference in TV signals which is most annoying at repetition rates around 5 cps.[7] Well-balanced video amplifier stages minimize sensitivity to additive components; cathode feedback and high-current operation of electron tubes minimize the nonlinear generation of bobble modulation components. However, to keep flicker interference within acceptable limits, the electronically regulated power supplies are counted on to reduce the effects of power line bobble by 50 db to 60 db. Electronic regulation plus dc operation of electron tube heaters and adequate filtering of unregulated power supplies control the amount of "hum" interference.

As a final result, the weighted interference to television by power line and other low-frequency noise in a terminal pair is expected to be 80 db below the peak-to-peak TV signal.

7.6 *Fluctuation Noise*

FM noise originating in the FM transmitter predominates over all other sources of fluctuation noise in FM terminals by at least 10 db. This noise originates in the two klystrons where it is generated by random (shot) processes in the electron beam.

Allowing for the frequency weighting of the interfering effect, for the action of typical TV circuit clampers and for the 12-db pre-emphasis, the klystrons introduce a fluctuation noise having the typical relative interference effect shown on Fig. 10(d). The integrated value of rms fluctuation noise in a terminal pair is 81 db below the peak-to-peak TV signal amplitude (8 mc peak-to-peak).

The relation of FM terminal noise to the over-all system noise for the telephone signal is discussed fully in Ref. 1. Briefly, noise interference

introduced by terminals is most serious in the telephone master-group at the lower end of the baseband. In a 3-kc band the rms frequency deviation due to noise is typically 35 cycles. Since an rms frequency deviation of 2.828 mc produces +8 dbm at the receiver output, the noise at this point is −90 dbm. At this point, the transmission level for the lowest level mastergroup is −25.5 db. Therefore, the noise is −64.5 dbm at 0 db TL; this is 17.5 dba in these "noisiest" telephone channels. If these channels remained at the same low-level end of the band (no frogging) for 16 terminals connected in tandem, the noise would add randomly to give a net noise meter reading of 29.5 dba at 0 db TL.

7.7 *Gain Stability*

Electronic stabilization of the most critical dc power supplies has reduced to insignificance the gain variations from these sources. However, power line voltage variation of ±1 per cent (typical stability) will cause ±0.2-db variation in the sensitivity of an FM receiver due to unregulated heater supplies. This random variation, added to systematic gain variations (estimated below as less than 0.05 db in one month), supports the expectation that gain changes of less than ±0.3 db will occur when an FM transmitter or an FM receiver is replaced by a standby unit through protection switching action in the field.

Presently available data on gain stability with time are meager, being the result of experimental observations over several month intervals on terminals in which the ages of all electron tubes were considerably under their life expectancy. Nevertheless, the performance under these condigave a reasonable degree of confidence in estimating that in a one month interval a typical terminal pair will have a systematic gain variation of −0.05 db and random variations not exceeding ±0.2 db. Sixteen terminal pairs in tandem would then have a net gain variation of −0.8 ± 0.8 db in a one-month interval.

APPENDIX A

Linearity, Differential Gain, Harmonic Distortion

Analytical expressions relating harmonic distortion to linearity and differential gain are obtained through the coefficients in the power series expansion for an output quantity (voltage, frequency deviation, etc.) as a function of input quantity. For example, consider the output vs. input characteristic

$$v_0 = a_1v + a_2v^2 + a_3v^3 + \cdots, \tag{1}$$

where the input, v, consists of a steady-state shift (or dc component), δ, added to a sinusoidal variation, $E \cos pt$; i.e., $v = \delta + E \cos pt$. Then, omitting dc terms, the output will contain the following sinusoidal components:

$$v_0 = F_1 \cos pt + F_2 \cos 2\,pt + F_3 \cos 3\,pt + \cdots$$

where

$$\begin{aligned} F_1 &= E[a_1 + 2a_2\delta + 3a_3\delta^2 + \cdots] \\ F_2 &= E^2\left[\frac{a_2}{2} + \frac{3}{2}a_3\delta + \cdots\right] \\ F_3 &= E^3\left[\frac{a_3}{4} + \cdots\right]. \end{aligned} \tag{2}$$

Differential gain is defined as the ratio of the small-signal gain (at fundamental frequency) for any shift, δ, to the gain when $\delta = 0$. Differential gain, as a function of δ, then has the form

$$\text{Differential gain} = 1 + L_1\delta + L_2\delta^2, \tag{3}$$

where, from the expression for F_1 in equation (2),

$$L_1 = \frac{2a_2}{a_1} \quad \text{and} \quad L_2 = \frac{3a_3}{a_1}.$$

Differential gain is normally expressed in db as $20 \log (1 + L_1\delta + L_2\delta^2)$.

Nonlinearity is closely related to differential gain and is defined as

$$\text{First-order nonlinearity} = L_1\delta \times 100 \text{ per cent} \tag{4}$$

$$\text{Second-order nonlinearity} = L_2\delta^2 \times 100 \text{ per cent}. \tag{5}$$

Harmonic performance can be expressed either in terms of the power series coefficients in (1) or the linearity coefficient in (3) as shown below:

$$\frac{\text{Second harmonic}}{\text{Fundamental}} = \frac{F_2}{F_1} = E\left[\frac{a_2}{2a_1} + \frac{3a_3}{2a_1}\delta - \left(\frac{a_2}{a_1}\right)^2 \delta \cdots\right]$$

The third term above is frequently negligible, in which case

$$\frac{F_2}{F_1} = E\left[\frac{1}{4}L_1 + \frac{1}{2}L_2\delta \cdots\right] \tag{6}$$

$$\begin{aligned} \frac{\text{Third harmonic}}{\text{Fundamental}} = \frac{F_3}{F_1} &= E^2\left[\frac{a_3}{4a_1} + \cdots\right] \\ &= E^2\left[\frac{1}{12}L_2 \cdots\right]. \end{aligned} \tag{7}$$

APPENDIX B

Klystron Quasi-Stationary Frequency Behavior

This appendix develops analytical relationships defining the FM properties of the 450A reflex klystron.

B.1 *Linearity of Steady-State Frequency Shift*

The detailed analysis of the behavior of reflex klystrons given in Ref. 10 suggests that a circuit analogue like that shown in Fig. 11 can be used to explain steady-state frequency behavior.

The phase shift, θ, in the drift space is a function of repeller voltage, v, and oscillating frequency, ω, while the phase shift, φ, in the resonator is a function of frequency only. To have sustained oscillations at any

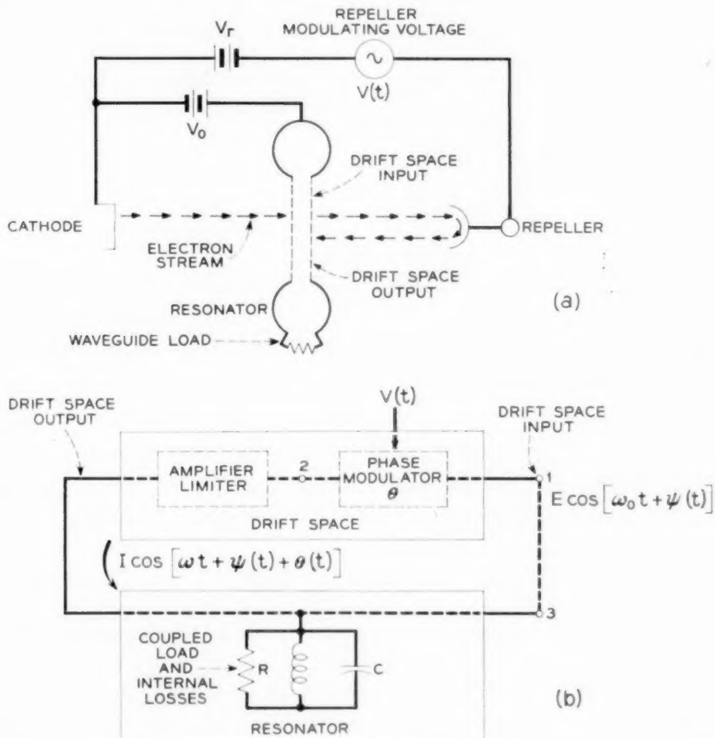

Fig. 11 — Circuit diagrams of reflex klystron: (a) schematic, (b) feedback loop.

frequency, the gain around the oscillator loop must be unity with a phase shift of zero. The equivalent amplifier-limiter action of the drift space takes care of satisfying the unity gain requirement. To satisfy the zero phase shift requirement, the oscillation frequency stabilizes at a value such that the resonator phase shift is equal and opposite to the drift space phase; i.e., $\varphi + \theta = 0$. If a change is made in the repeller voltage to $v + dv$ such that the delay in the drift space increases and the drift space phase changes to $\theta + d\theta$, then the oscillating frequency will decrease sufficiently to give an equal and opposite phase change in the resonator. Nonlinearity in the frequency-voltage relationship results primarily from the nonlinear relationship between frequency and phase in the resonator and to a lesser extent because the changes in drift space delay are not linearly related to changes in the repeller voltage. These relationships will be evident from solutions to the differential equation:

$$d\varphi + d\theta = 0 \tag{8}$$

from which

$$\frac{d\varphi}{d\omega} d\omega + \frac{\partial\theta}{\partial v} dv + \frac{\partial\theta}{\partial\omega} d\omega = 0 \tag{9}$$

and

$$\frac{d\omega}{dv} = -\frac{\dfrac{\partial\theta}{\partial v}}{\dfrac{d\varphi}{d\omega} + \dfrac{\partial\theta}{\partial\omega}}. \tag{10}$$

In Ref. 10, p. 593, the relationship between drift space phase and electrode voltages is given as

$$\theta = \frac{k\omega\sqrt{V_0}}{V_0 + V_r} \tag{11}$$

where the voltages V_0 and V_r are those shown in Fig. 11. Also, assuming that the resonator behaves as a simple parallel resonant circuit,

$$\varphi = \text{arc tan } Q\left[\frac{\omega}{\omega_r} - \frac{\omega_r}{\omega}\right]. \tag{12}$$

These relationships can be differentiated and the resulting expressions expanded in power series which are valid in the vicinity of the operating point of the klystron. When the first few terms of these series are substituted in (10), an expression for the small-signal FM sensitivity as a function of the frequency shift, $\delta = \omega - \omega_r$, and repeller voltage,

V_r, is obtained. The repeller voltage can be eliminated as a variable in the expression by substituting $V_r = V_{r0} + a\delta$, where $a = dv/d\omega$ evaluated at $\omega = \omega_r$ and V_{r0} is the repeller voltage corresponding to an oscillating frequency ω_r. The first few terms then become

$$\frac{d\omega}{dv} \approx \frac{\omega_r}{(V_0 + V_{r0})} \frac{1}{\left(\frac{2Q}{\theta_0} + 1\right)} \left[1 - \left(\frac{4Q}{\theta_0} - 1\right)\frac{\delta}{\omega_r} + \frac{8Q^3}{2Q + \theta_0}\frac{\delta^2}{\omega_r^2}\right], \quad (13)$$

where

$\delta = \omega - \omega_r$,

ω_r = resonant frequency of cavity in radians/sec,

θ_0 = drift space phase angle in radians, at ω_r,

Q = cavity Q, and

$V_0 + V_{r0}$ = dc potentials applied to klystron as shown in Fig. 11.

Comparing this expression with (3) we write

$$\text{Differential sensitivity} = 1 + L_1\delta + L_2\delta^2 \quad (14)$$

where

$$L_1 = -\left(\frac{4Q}{\theta_0} - 1\right) \Big/ \omega_r,$$

and

$$L_2 = \frac{8Q^3}{(2Q + \theta_0)\omega_r^2}.$$

Typical numerical values for these quantities are:

$\omega_r = 2\pi \times 6 \times 10^9$,

$\theta_0 = 2\frac{3}{4}$ cycles = 17.3 radians,

Q = 100, and

$V_0 + V_r = 550$;

which give:

$$\text{Differential sensitivity} = 1 - 22\frac{\delta}{\omega_r} + 37{,}000\,\frac{\delta^2}{\omega_r}.$$

For a ±4-mc frequency deviation ($\delta/\omega_r = 4/6000$), the first-order non-

linearity is ± 1.5 per cent, and the second-order nonlinearity is 1.6 per cent. The first-order nonlinearity is primarily due to the nonlinear relationship between drift space phase shift and repeller voltage, while the second-order nonlinearity is predominently caused by the nonlinear phase shift in the resonator as a function of frequency.

B.2 *Second Harmonic Balance*

It is evident from the expression for F_2/F_1 in (6) that the second harmonic can be balanced to zero by selection of the operating point δ, such that $\delta = \delta_0 = -L_1/2L_2$. In terms of (14) a second harmonic balance can be obtained by means of a small change in the klystron repeller voltage so that the quiescent oscillating frequency is slightly different from the resonant frequency of the cavity, the amount of this difference being given by

$$\frac{\delta_0}{\omega_r} \approx \frac{\dfrac{4}{\theta_0} - 1}{\dfrac{16Q^3}{2Q + \theta_0}} = 3.0 \times 10^{-4}.$$

This corresponds to shifting to an operating point which is approximately 1.8 mc above the resonant frequency of the cavity.

B.3 *Experimental Adjustment for Second Harmonic Balance*

As can be checked by differentiating (13) and setting the resulting expression equal to zero, *minimum* differential sensitivity occurs at the same value of δ as was found for zero second harmonic. This principle is employed in a field adjustment procedure for "optimizing" klystron linearity: repeller bias is adjusted to minimize the small-signal FM deviation produced by a small voltage variation.

B.4 *Second-Order Balance Stability*

As a Function of Repeller Voltage. It has been shown above that it is possible to adjust the klystron bias so as to obtain a second harmonic balance. For such an adjustment to be worthwhile, however, the operating point must stay within bounds. For example, the ratio of the second harmonic to fundamental is given in (6) as

$$\frac{F_2}{F_1} = E\left(\frac{1}{4} L_1 + \frac{1}{2} L_2\delta\right),$$

where for $E = 4$ mc, the requirement on F_2/F_1 is -54 db* or 0.002. Substitution of the numerical values above gives a requirement on the frequency shift of 1.8 ± 1.0 mc. With a repeller sensitivity of about 1.45 mc per volt, the voltage stability requirement is ± 0.7 volt for the -100-volt repeller supply.

As a Function of the Resonator Voltage. The resonator has a sensitivity of approximately 0.5 mc per volt. Therefore, the frequency stability requirement of $+1.0$ mc, just derived, imposes a voltage requirement of about ± 2.0 volts on the 450-volt resonator supply. Thus, regulated supplies are required for both the repeller and resonator to maintain the required second-harmonic balance.

As a Function of Resonant Frequency. If temperature or tuning changes the resonant frequency of the cavity, the phase shift in the cavity is a function of two variables: the oscillating frequency, ω_0, and the resonant frequency, ω_r. With V constant, (9) becomes

$$\frac{\partial \varphi}{\partial \omega_r} d\omega_r + \frac{\partial \varphi}{\partial \omega_0} d\omega_0 + \frac{d\theta}{d\omega_0} d\omega_0 = 0$$

where

$$\frac{\partial \varphi}{\partial \omega_r} = -\frac{\partial \varphi}{\partial \omega_0}$$

for $\omega_0 \approx \omega_r$

Substituting, and solving for $d\omega_0/d\omega_r$ gives

$$\frac{d\omega_0}{d\omega_r} = \frac{\dfrac{\partial \varphi}{\partial \omega_0}}{\dfrac{\partial \varphi}{\partial \omega_0} + \dfrac{d\theta}{d\omega_0}}.$$

For the 450A klystron

$$\partial \varphi/\partial \omega_0 \approx \frac{2Q}{\omega_r} \approx 5.3 \times 10^{-9} \text{ second}$$

$$d\theta/d\omega_0 \approx \frac{\theta_0}{\omega_0} \approx 0.46 \times 10^{-9} \text{ second.}$$

Therefore,

$$\frac{d\omega_0}{d\omega_r} \approx 0.92$$

* Only a portion of the total requirement of -49 db is allocated to the klystron.

and as the resonant frequency changes, the oscillating frequency tends to follow so that the change in the desired 1.8-mc offset is only 8 per cent of the change in the resonant frequency.

Thermal and mechanical stabilities in the 450A klystron are such that second-order balance degradation due to rest frequency shifts caused by resonant frequency changes are generally negligible compared to the degradation due to resonator and repeller supply voltage changes.

B.5 *Third-Harmonic Distortion Performance*

The low-frequency third-harmonic performance of the 450A klystron with a resonator $Q = 100$ and for a peak frequency deviation of 4 mc is given by (7) with $E = 4$ mc and

$$L = \frac{8Q^3}{(2Q + \theta_0)\omega_r^2}.$$

Thus,

$$\frac{\text{Third harmonic}}{\text{Fundamental}} = E^2 \times \frac{8Q^3}{12\omega_r^2(2Q + \theta_0)}$$

$$= 0.00137 \text{ or } -57.2\ db.$$

This is 6 db better than the objective of -51 db.

APPENDIX C

Klystron Dynamic Behavior—Video Response

It was tacitly assumed in Appendix B that the steady-state frequency deviation behavior is also applicable when the incremental repeller voltage is a function of time. This assumption has been verified experimentally; frequency deviation is substantially independent of video signal frequency up to at least 10 mc. Similarly, harmonic distortion depends upon Q as predicted by the steady-state analysis.

This dynamic behavior may also be verified theoretically. The essence of this theoretical analysis will be outlined here to show how the integrating action of the resonator dynamically converts phase modulation to frequency modulation. Appendix D shows how this same action converts noise in the electron stream to an FM deviation noise in the output.

When a signal voltage is applied to the repeller in Fig. 11(b), a phase modulation of the carrier is produced in the closed loop. Finding a relationship between the phase modulation (output) and the modulating signal (input) is analogous to the problem encountered in any closed-loop

system such as a feedback amplifier. In principle, the loop between (1) and (3) on Fig. 11(b) is temporarily opened, and an arbitrarily amplitude and phase modulated signal,

$$e(t) = E_1(t) \cos [\omega_0 t + \psi(t)], \tag{15}$$

is introduced at 1. This signal is traced around the loop and appropriately modified until the output at 3 is expressed in terms of the arbitrary wave introduced at 1. At this point the input and output waves for the open loop can be equated to determine the closed-loop performance of the system. Since the output signal, as well as the input signal, will have the form of an amplitude and phase modulated wave, the amplitude modulation and phase modulation can be separately equated. Only the results for the phase modulation terms are of interest here. Some simplification is achieved by working in the frequency domain. Thus, the following definitions are made.

$S_\psi(\omega)$ = frequency spectrum of $\psi(t)$, and

$S_\theta(\omega)$ = frequency spectrum of $\theta(t)$, the phase modulation introduced by the modulating signal, $V(t)$.

The spectrum of the phase modulation at the output of the drift space becomes

$$S_\psi(\omega)e^{-j\omega D} + S_\theta(\omega)$$

where D equals the delay in drift space in seconds.

This spectrum is again modified as it passes through the resonator. The resonator has the effect of a low-pass filter on the phase modulation, where the filter is the low-pass equivalent of the actual bandpass structure. It thus consists of a resistor in parallel with a capacitor with values such that the bandwidth of the low-pass structure is just half that of the actual bandpass structure. For a Q of 100 at 6000 mc, the bandwidth is about 60 mc; the equivalent low-pass bandwidth is 30 mc, from which the low-pass transmission characteristic is therefore

$$Y(\omega) = \frac{1}{1 + j\omega\tau}, \tag{16}$$

with

$$\tau = 2Q/\omega_0 = 5.3 \times 10^{-9} \text{ second.}$$

The spectrum of the phase modulation at the output of the open loop is therefore

$$Y(\omega)S_\psi(\omega)e^{-j\omega D} + Y(\omega)S_\theta(\omega)$$

and under closed loop conditions

$$S_\psi(\omega) = Y(\omega)S_\psi(\omega)e^{-j\omega D} + Y(\omega)S_\theta(\omega) \tag{17}$$

from which

$$S_\psi(\omega) = \frac{Y(\omega)}{1 - Y(\omega)e^{-j\omega D}} S_\theta(\omega). \tag{18}$$

Substitution for $Y(\omega)$ as given in equation (16) permits the excellent approximation,

$$S_\psi(\omega) \approx \frac{1}{j\omega(\tau + D)} S_\theta(\omega) \tag{19}$$

which for the 450A klystron is accurate to within about 0.01 db up to the top baseband frequency of 10 mc. After both sides of (19) are multiplied by $j\omega$, the following identifications are made:

$j\omega S_\psi(\omega)$ = frequency spectrum of $\psi'(t)$, the first time derivative of $\psi(t)$, and

$S_\theta(\omega)$ = frequency spectrum of $\theta(t)$.

Since the instantaneous frequency of a wave is defined as the first time derivative of the instantaneous phase, the time domain equivalent of (19) can be written as

$$\text{frequency modulation} = \psi'(t) = \frac{\theta(t)}{\tau + D}. \tag{20}$$

Thus, within the accuracy of the approximation stated above, the closed-loop frequency modulation is directly proportional to the phase modulation introduced by the modulating signal. Furthermore, the equivalence of this result to the quasi-stationary result obtained in Appendix B is possible when the following identifications are made:

$$\tau = \frac{2Q}{\omega_r} = \frac{d\varphi}{d\omega}$$

$$D = \frac{\partial\theta}{\partial\omega}.$$

Thus, the above result can also be written in the form

$$\varphi'(t) = \frac{\theta(t)}{\dfrac{d\varphi}{d\omega} + \dfrac{\partial\theta}{\partial\omega}} \approx \frac{\dfrac{\partial\theta}{\partial v}}{\dfrac{d\varphi}{d\omega} + \dfrac{\partial\theta}{\partial\omega}} v(t)$$

which is the linearized dynamic equivalent of (10).

From the preceding analysis the following conclusions are obtained:

(a) Phase modulation is dynamically converted to frequency modulation by regenerative action in the feedback loop.

(b) Klystron modulation sensitivity, inversely proportional to open loop delay, is the same as that found in the steady-state analysis.

(c) The resultant frequency deviation is essentially independent of video signal frequency.

APPENDIX D

Klystron Fluctuation noise

A comparison of the FM noise characteristic of 450A klystrons, Fig. 10(d), with noise characteristics typical of other electron tube devices leads to the conclusion that the two phenomena are identifiable. "Flicker" noise, varying approximately as $1/f$, predominates for frequencies less than 25 kc. "Shot" noise, flat with frequency, predominates at frequencies greater than 25 kc. Since shot noise is controlling over all of the message band and a major portion of the television band, it is the more important source of klystron noise.

Flicker noise is associated with random time variations in group electron emission from the cathode. The power spectrum for these variations decreases at about a $1/f$ rate. The exact mechanism whereby this random amplitude modulation of the dc space current is converted to frequency modulation of the carrier has not been quantitatively identified. It probably comes from variations in the inter-action grid capacitance or drift space delay induced by the space charge density modulation.

On the other hand, shot noise has been clearly identified as being due to the statistical time variations for single electron emissionor (capture by intervening grids), which have spectral intensities that are relatively constant up to and above 6000 mc. The electron stream flowing through the resonator interaction grids [twice, as illustrated in Fig. 11(a)] varies randomly with time. Those frequency components of the variations which fall in bands equally displaced on either side of the carrier add to it, and randomly vary its phase. This phase variation is converted to random frequency deviation by the action of the regenerative loop as described in the previous section.

Shot noise per cycle of bandwidth is given by the following equation for temperature-limited emission:

$$i_s^2 = 2(3.18 \times 10^{-19})\, I \tag{21}$$

where i_s is the rms fluctuation current and I the dc beam current. The

factor 2 is used to allow for the double transit of the electron stream through the interaction grids.

Also flowing into the interaction grid space is a bunched electron stream which has an equivalent fundamental carrier current component, i_c, at radian frequency, ω_0. This current produces power, P, in the resonator load. Therefore,

$$i_c = \sqrt{\frac{P}{R}} = \sqrt{\frac{P\omega_0 C}{Q}} \tag{22}$$

where

R = resistance of the loaded resonator, and

C = capacitance of the interaction grid space.

This carrier current will be randomly phase modulated by upper and lower sideband noise currents to give a net open-loop rms phase deviation in each one cps band of

$$\theta_n = \frac{i_s}{i_c} = i_s \sqrt{\frac{Q}{P\omega_0 C}} \text{ radians}. \tag{23}$$

Regenerative action described in Appendix C will change this to a frequency deviation which is independent of the noise spectrum frequency. Neglecting D with respect to τ in equation (20):

$$\begin{aligned} \text{rms frequency deviation} &= \frac{\theta_n}{\tau} = \frac{\theta_n \omega_0}{2Q} \\ &= i_s \sqrt{\frac{\omega_0}{4PQC}} \text{ rad/sec in a one-cps band due to noise.} \end{aligned} \tag{24}$$

Values typical of the 450A klystron are:

$2I$ = 0.1 amp,

P = 0.2 watt,

C = 0.5 micromicrofarad,

Q = 100, and

$\omega_0 = 2\pi \times 6175 \times 10^6$.

When substituted in (24), these give an rms frequency deviation in a one-cps band due to shot noise of 0.9 cps.

Experimentally, it is found that one 450A klystron generates an rms

frequency deviation of about 0.4 cps in a one-cps band*; for the two klystrons in the FM modulator the noise is 1.41 times greater.

Apparently space-charge smoothing, which usually reduces rms shot noise in space-charge-limited devices by a factor of 5, gives a 2/1 improvement in this klystron. Alternatively, space-charge smoothing may be more fully effective, and partition capture may be responsible for the added noise.

APPENDIX E

Discriminator Network Design

To demonstrate analytically the main features of the discriminator network design, the magnitudes of the interstage impedances for the common interstage and for the high- and low-frequency discriminator networks as shown in Fig. 7 are expanded in power series about the center frequency of the discriminator. For the common interstage this gives

$$| Z_c | = | Z_{c0} | (1 + c_1\delta + c_2\delta^2 + c_3\delta^3 \cdots), \tag{25}$$

where

$$\delta = \omega - \omega_0 \text{, and}$$

$$\omega_0 = \text{center frequency of the discriminator.}$$

Similarly, for the high- and low-frequency discriminator networks,

$$| Z_l | = | Z_{l0} | (1 + l_1\delta + l_2\delta^2 + l_3\delta^3 + \cdots) \tag{26}$$

$$| Z_h | = | Z_{h0} | (1 + h_1\delta + h_2\delta^2 + h_3\delta^3 + \cdots). \tag{27}$$

The amplitude characteristics for transmission through the two sides of the discriminator are then obtained as the product of $| Z_c |$ with $| Z_l |$ and $| Z_h |$ respectively. Thus,

$$A_l = k_l | Z_c | | Z_l | \tag{28}$$

$$A_h = k_h | Z_c | | Z_h | \tag{29}$$

where the factors k_l and k_h are introduced to include the effect of electron tube gains in the two sides.

For the purpose of the following discussion the constant, c_1, will be taken equal to zero. This is done in the actual design by tuning the in-

* To generate this same noise by thermal agitation would require a resistance in the repeller circuit of 5 megohms; the actual circuit value is 800 ohms.

terstage network to the center frequency of the discriminator. Performing the multiplications indicated in (28) and (29) and discarding terms higher than δ^3 gives

$$A_l = A_{l0}[1 + l_1\delta + (l_2 + c_2)\delta^2 + (l_3 + c_3 + l_1c_2)\delta^3] \tag{30}$$

$$A_h = A_{h0}[1 + h_1\delta + (h_2 + c_2)\delta^2 + (h_3 + c_3 + h_1c_2)\delta^3] \tag{31}$$

where

$$A_{l0} = k_l \mid Z_{c0} \mid \mid Z_{l0} \mid \tag{32}$$

$$A_{h0} = k_h \mid Z_{c0} \mid \mid Z_{h0} \mid . \tag{33}$$

The input signal has the form,

$$i(t) = i_0[1 + A(t)] \cos [\omega_0 t + \varphi(t)] \tag{34}$$

where

$$A(t) = \text{amplitude modulation,}$$

$$\varphi(t) = \text{phase modulation, and}$$

$$\varphi'(t) = \text{frequency modulation.}$$

As a result of the amplitude shape in the discriminator networks, the instantaneous amplitudes at the output of the networks (or input to AM detectors) are approximately* as follows:

$$M_l(t) = i_0A_{l0}[1 + A(t)]$$
$$[1 + l_1\varphi' + (l_2 + c_2)\varphi'^2 + (l_3 + c_3 + l_1c_2)\varphi'^3] \tag{35}$$

$$M_h(t) = i_0A_{h0}[1 + A(t)]$$
$$[1 + h_1\varphi' + (h_2 + c_2)\varphi'^2 + (h_3 + c_3 + h_1c_2)\varphi'^3] \tag{36}$$

The output of the balanced discriminator is given by the difference between the outputs of the two detectors. If the diode detector efficiencies for the two sides of the discriminator are D_l and D_h, the discriminator output is given as

$$\text{Discriminator output} = D_hM_h(t) - D_lM_l(t). \tag{37}$$

* These results are based on a quasi-stationary approach and are valid for very low modulating frequencies. At higher modulating frequencies the relationship between instantaneous frequency and instantaneous amplitude is more complex. The additional terms which then arise are due primarily to delay distortion which is compensated by delay equalization (see Section 7.2).

The individual output terms are listed and discussed below. For simplicity in the expressions which have to be written we let $E_l = i_0 A_{l0} D_l$ and $E_h = i_0 A_{h0} D_h$.

(i) $E_h - E_l =$ dc output.

(ii) $(E_h - E_l)A(t) =$ output due to unsuppressed amplitude modulation. By adjusting the relative gains in two sides of the discriminator so that the dc output is zero, this term is also eliminated.

(iii) $(E_h h_1 - E_l l_1)\varphi'(t) =$ desired signal output. Since h_1 is positive and l_1 is negative, the signal components from the two sides actually add in the output.

(iv) $[E_h(h_2 + c_2) - E_l(l_2 + c_2)]\varphi'(t)^2 =$ unwanted second-order modulation. This term will be discussed in detail later.

(v) $[E_h(h_3 + c_3 + h_1 c_2) - E_l(l_3 + c_3 + l_1 c_2)]\varphi'(t)^3 =$ unwanted third-order modulation.

(vi) Finally there is a set of terms identical to (iii), (iv), and (v) above except that each is multiplied by $A(t)$. By keeping $A(t)$ small by means of limiter action ahead of the discriminator, these terms are held to acceptable levels.

The effect of the common interstage is demonstrated by examination of the distortion term listed in (iv) above. In the absence of the common interstage, this term would be

$$[E_h h_2 - E_l l_2]\varphi'(t)^2$$

and, with $E_h = E_l$ to minimize distortion term (ii), it is desirable, and possible, to adjust the discriminator so that $h_2 = l_2$ and the term goes to zero. For the values of h_2 and l_2 of the actual discriminator design, the requirement on the stability of E_h and E_l to meet the modulation objective would then be given by

$$|E_h - E_l| \leqq \frac{|E_h|}{15}.$$

This corresponds to holding the relative gains of the two sides of the discriminator equal to within about 0.5 db between maintenance intervals.

The use of a common interstage selected so that $c_2 = -h_2 = -l_2$ causes the distortion term to go to zero even though E_h and E_l are not equal. This is illustrated in Fig. 9(a). In practice, it has been found

possible to adjust c_2 such that $c_2 + h_2 = c_2 + l_2$ are about 10 per cent of $h_2 = l_2$, which relaxes the requirement on gain stability to

$$| E_h - E_l | \leqq \frac{| E_h |}{1.5},$$

which corresponds to a relative gain stability of 4.4 db.

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Power Systems for the TH Radio System

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The basic source of power for the TH radio equipment is reliable 230 volts ac obtained from a bank of special motor-alternator sets. These are normally driven by commercial ac power but switch automatically to battery drive on loss of commercial power. This firm ac power is distributed to the individual units of equipment where local rectifiers, of various types and sizes, produce dc power of the required voltage and stability. One of the more advanced of these is the 2900-volt regulated supply for two traveling-wave tubes.

I. GENERAL

The transmission and control equipments in the TH radio system require dc power at a multiplicity of voltages from 7 to 3100 volts. A broad range of stability, noise and other performance requirements is established according to the sensitivities of the individual loads. High reliability is an objective common to all of the power equipment. Other important objectives are flexibility for growth, simplicity for maintenance, protection of personnel from hazardous voltages, and minimum cost.

1.1 *Characterization of Loads*

The dc loads can be placed in four general categories:

(i) Low to medium voltages (7 to 250 volts), characterized by constant load, tolerance to normal voltage variations of ± 2 per cent, and noise or ripple requirements that can be satisfied economically with passive filters.

(ii) High voltages (1200 to 3100 volts) for the traveling-wave tubes, characterized by $\pm \frac{1}{2}$ per cent stability requirements and several special control and protection features.

(iii) Medium voltages (-170, $+220$, and $+450$ volts) for the klys-

trons and video amplifiers in the FM terminal equipment, characterized by ± 0.2 per cent long-term stability requirements, and noise requirements as low as 0.0001 per cent in the frequency range of $\frac{1}{2}$ to 30 cps where passive filtering is impractical.

(iv) $+24$ and -24 volts for the protection switching equipment with especially stringent reliability requirements.

1.2 *Design Approach*

While many of the voltages (12, 24, 130, and 250) could be supplied directly from standard battery plants, the traveling-wave tube voltages are too high for this type of approach. Instead, special motor-alternator sets are employed to generate reliable and continuous (firm) ac power. The motor-alternators are normally driven by commercial ac power but switch over to battery drive in the event of a commercial power outage.

All dc voltages are produced by dc power supplies connected to one of the firm ac busses. All dc power supplies are decentralized such that a single failure can affect no more than a single one-way broadband channel. (An exception to this principle is the supply to the protection switching equipment where redundant rectifiers and direct battery reserve are provided.) Decentralization of the dc power supplies simplifies the distribution of power and provides flexibility for growth.

II. PRIME AND STANDBY POWER SOURCES

Commercial service is considered the normal prime source of power for a TH radio station. Under commercial power failure conditions, emergency power is provided by single or multiple diesel engine-alternator sets. Power service for the TH system has been standardized at three-phase, four-wire 120/208-volt input because of high power requirements. The ac distribution system for a typical installation, including the firm ac motor-alternator arrangements described in Section III, is shown in Fig. 1.

Commercial ac loads are divided into two separate categories, namely, protected and unprotected. Protected loads are those which require emergency service during commercial power failure conditions. Unprotected, or nonessential, equipment loads are supplied from a separate distribution cabinet and include such items as normal office lighting, air conditioning and the motor-generator or rectifier used for recharging the 130-volt battery. Protected, or essential, equipment loads form, by far, the larger percentage of the total office equipment, as may be seen from Fig. 1. The prime subdivision of these loads is accomplished through

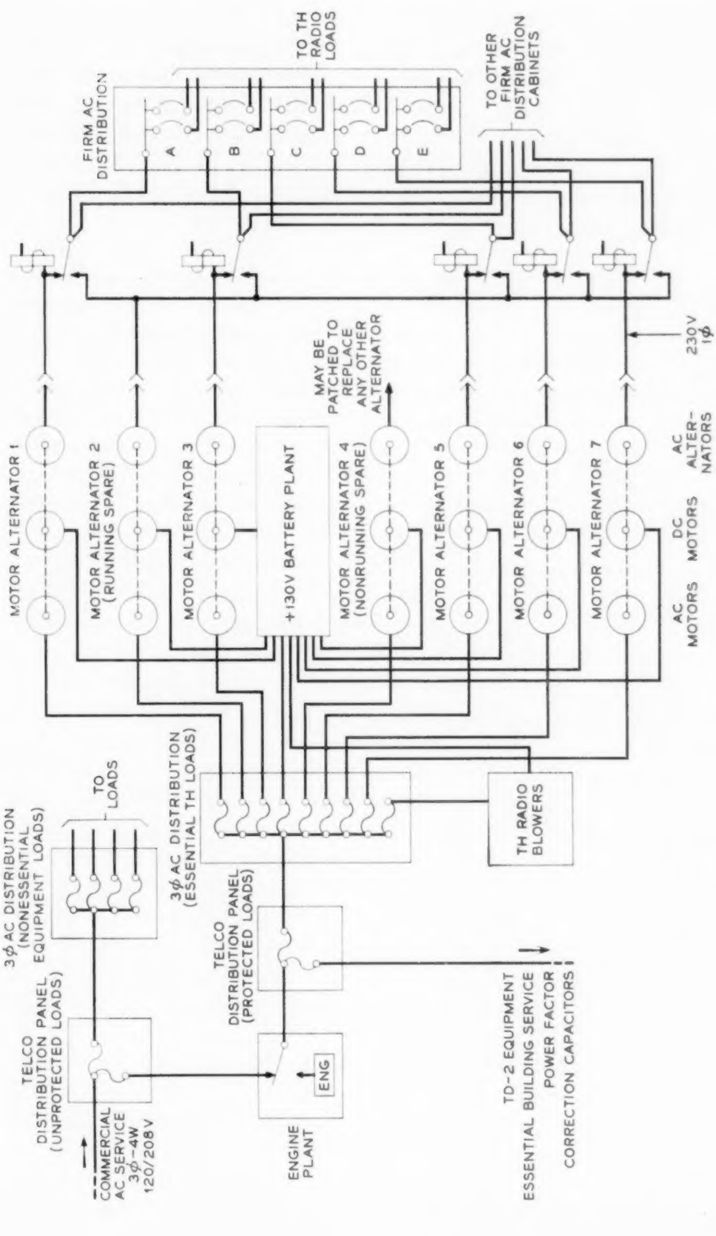

Fig. 1 — AC power distribution for typical TH installation.

a customer-furnished distribution cabinet. This cabinet serves all essential building services and power factor correction capacitors when required to obtain maximum engine-alternator capacity, and is the main feed when TH systems are combined with TD-2 stations. All TH loads are derived from this cabinet via an intermediate three-phase distribution cabinet. This latter unit distributes power to essential loads such as the 508A motor-alternator plant, the 130-volt battery plant, the radio tube cooling system, tower lighting, utility outlets, etc.

III. FIRM AC POWER GENERATION

The firm ac power system is designed for completely automatic unattended operation at remote and, at times, inaccessible mountain locations. This supply is provided by a two-motor generator set which couples together on one shaft a synchronous induction ac motor, a dc motor and a brushless ac type alternator. Under normal operating conditions the set is started manually by its dc motor from the station 130-volt battery. After reaching normal speed and voltage, the set is placed in normal automatic operation by connecting it to the load and transferring to ac motor drive. With normal line frequency, synchronous motor drive, and essentially fixed load, the alternator output voltage is constant and insulated from line voltage transients and variations. Line voltage monitors, set above the pull-out torque point of the ac motors, control automatic transfer to dc motor drive in the event of marginal line voltage and permit return to ac drive when the ac input is restored to its normal voltage and frequency range. Dc operation for about three minutes after power returns gives the line a chance to stabilize before permitting ac drive. If the line frequency is outside allowable limits, causing abnormal alternator speeds and, thereby, abnormal load voltages, monitors on each alternator output control automatic transfer to dc drive until the supply frequency returns to its normal range. If the output of any alternator goes beyond allowable limits for other than excessive supply frequency variations (e.g., an individual ac motor failure), then the output voltage monitor controls automatic transfer and lockover to dc motor drive with alarms to indicate a trouble condition on a particular alternator.

The output monitor on each alternator also guards against a continued abnormal voltage if the output drops below 90 per cent of the normal 230-volt value for more than 30 milliseconds or below 95 per cent for more than three seconds. Under these unlikely conditions, the control circuits transfer the load to the running spare, shut down the alternator involved, and cause an alarm to be given. As shown in Fig. 1, alternator

2 is a common running spare for all the regular alternators. It takes over the load of any regular alternator which fails or which is manually shut down for maintenance. Should more than one alternator fail at one time, then the lowest numbered alternator gets preference on the basis that the most important loads are allocated to the first alternators. A minimum installation requires four alternators serving the two busses A and B. This involves regular alternators 1 and 3, hot running spare alternator 2, and cold spare alternator 4. Added loads on busses C, D and E require another regular alternator for each bus. Alternator 4 is for use on a plug-in basis to replace any other alternator for major repairs. It uses the output control circuits of the alternator it replaces but has its own motor input circuits.

3.1 *Two-Motor Alternator Design*

A new type of two-motor alternator set has been developed for the coded 508A plant for this application. It is designed for normal brushless operation; a rotating field exciter using rotor mounted rectifying diodes for the alternator field self excitation eliminates the heretofore conventional use of a shunt generator exciter.

The set has a new design of induction motor to operate at synchronous speeds instead of at slip speeds as with the ordinary squirrel cage induction motor. The synchronous characteristics of this new type of motor are produced by the special construction of a die cast aluminum rotor. This not only has salient poles, as in a reluctance type synchronous motor, but also has internal flux guiding paths which exert many times the synchronizing torque of a reluctance synchronous motor. The motor acts like an ordinary induction motor, accelerating until it almost reaches synchronism, whereupon the synchronizing torque pulls the rotor into synchronism with the magnetic field of the stator. These motors give synchronous operation without rotor windings and without slip rings and brushes for dc excitation.

In keeping with the brushless ac motor, the dc motor is operated normally with its brushes lifted by ac solenoids but with its field continuously excited from battery. Failure of ac will drop the brushes to pick up battery drive. Brush lifting has presented design problems involving proper brush travel and insulation to break 130-volt dc arcs safely, proper brush positioning and contact when released, and positive lifting when the solenoids are energized.

The dc motor speed characteristics are designed to work with an external regulating circuit to maintain speed within one cycle of normal 60-cycle speed for battery voltages between 145 and 120 volts. This

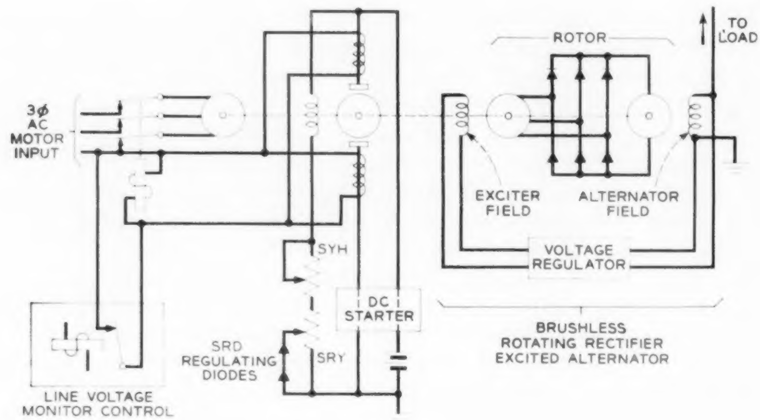

Fig. 2 — Circuit of the two-motor alternator set.

regulating circuit, shown in Fig. 2, utilizes the reverse characteristics of Zener diodes to provide a nonlinear shunt around fixed resistance in series with the shunt field. At the minimum battery voltage, 120 volts, and with the diodes nonconducting, the resistance in series with the field is adjusted by SYH to give 60 cycles at a speed of 1800 rpm. The setting of the SRY potentiometer is such that any increase in battery voltage above 120 volts increases the drop across this portion of the external field circuit, until the diodes break down to hold this drop constant. Battery voltages above 120 volts thus increase the field current faster than in direct proportion to the voltage rise. For example, without diodes the increase in field current would equal the ratio of 145/120 or 1.21, whereas with the diodes the ratio is increased to 1.34. The dc motor speed characteristic is designed so that this nonlinear increase in field current offsets the direct increase in armature current and thus maintains approximately constant speed over the battery discharge voltage range.

As an example of the dc motor speed regulation with varying battery voltage, Table I is taken from laboratory test data.

Table I — DC Motor Speed Regulation

Battery Volts	Field Volts	Diode Volts	Field Amps	Diode Amps	Alternator Freq
125	48.5	50	0.542	0.017	59.9 cps
145	60.5	51.3	0.678	0.678	60.1 cps

Continuing with the brushless theme, the set uses a recently developed type of alternator, which employs an ac generator winding with rotor-mounted rectifiers to furnish dc for the alternator field. A voltage regulator controls the direct current on the exciter field and maintains the load voltage within ± 1 per cent from no load to full load for power factors 0.7 to unity. The machine design, together with the fast acting transistor amplifier driven magnetic regulator, minimizes transient surges and voltage overshoots for load changes. The effect of transferring the alternator drive from ac to dc results in a $2\frac{1}{2}$ per cent voltage dip over a 100-millisecond interval until normal output voltage is restored.

Fig. 3 shows a typical power installation including the firm ac plant, the associated emergency 130-volt battery plant, and the commercial ac distribution cabinets.

The over-all alternator set is a single frame two-bearing unit with the dc motor at one end to give access to brushes and commutators and the alternator at the opposite end to provide ready access to the rotating rectifiers. The set mounts on the two-tier framework seen in Fig. 3. The associated control equipment is in the adjoining control bay cubicle. The bay is arranged for single side maintenance with swinging gate and hinged panels.

The equipment is completely coded in shop assembled and wired units ready for a minimum of connections in the field and requiring little job

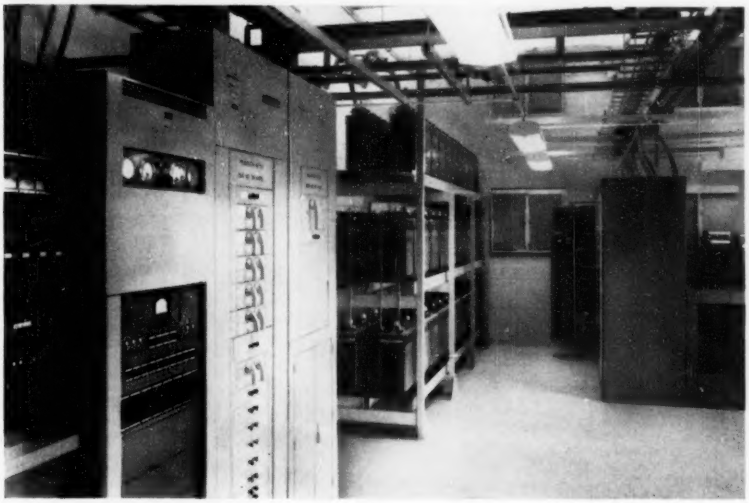

Fig. 3 — Typical power installation for the TH Radio System.

engineering for ordering initially and for additions. One equipment bay and framework are required for each two alternators for a maximum of four units for the ultimate of seven alternators.

3.2 *Firm AC Distribution*

The distribution arrangements for the firm ac power plant are engineered to utilize fully the capacity of the motor-alternators, not only initially but ultimately, and at the same time to realize the maximum in reliability for the working radio circuits. This requires considerable engineering and planning initially to anticipate future requirements and to avoid assignments which would make protection or standby facilities unavailable in the event of multiple failure in the motor-alternator plant resulting in loss of power on a firm ac distribution bus (A to E of Fig. 1). An ordinary repeater station, when fully equipped, has four normal running motor-alternators, each carrying two two-way broadband channesl; the microwave carrier supply and auxiliary channel equipment are evenly divided between busses A and B. For other types of stations, such as switching main stations, where auxiliary services like protection switching and terminal facilities are required, careful planning of ac assignments enables the service protection features to be met. Stations of this latter type require the ultimate of five normally running machines.

IV. POWER SUPPLIES

4.1 *Unregulated Rectifiers*

For general purpose use, a number of unregulated rectifiers were designed, as shown in Table II.

These rectifiers operate from the firm 230-volt ac and employ diffused silicon diodes and conventional passive filters. A limited voltage adjustment range is provided by a variable transformer bridged across a por-

TABLE II — UNREGULATED RECTIFIERS

dc Voltage	Current Rating	RMS Ripple	Applications
7 v	1 amp	0.07 v	Switch Control (Carrier Supply)
11	2	0.5	Heater and Bias
11	6	0.5	Heater and Bias
24	2	0.35	Switch Control (Carrier Supply)
48	1.5	0.5	ON Terminals (Aux. Chan.)
135	0.125	0.025	Plate Supply
135	0.325	0.025	Plate Supply
250	0.22	0.025	Some Plate Loads (Carrier Supply)

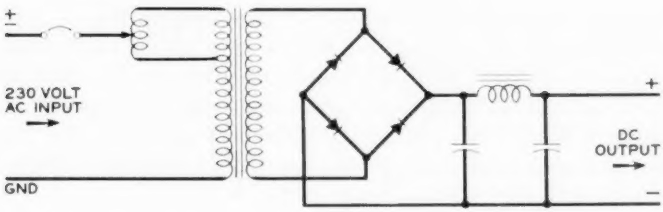

Fig. 4 — Typical circuit of adjustable, unregulated rectifier.

tion of the transformer primary winding. A typical circuit is illustrated in Fig. 4. The transient surge currents occurring in the diodes (capacitor charging current) and in the transformer primary winding (transformer inrush current plus capacitor charging current) were carefully investigated to avoid diode failure and nuisance circuit breaker operation. These transient investigations were aided by use of a special ac input switch which may be preset to close at a selected phase angle of the 60-cycle ac voltage wave.

All of these rectifiers are individual packages and bear a strong family resemblance. Several are shown in Fig. 5. Since the rectifiers are designed for minimum volume, there results a variety of sizes. They hang on vertical panels in the TH bays by means of keyhole slots in the bay panels which are engaged by captive screws in the underside of the rectifier chassis. Capacitors and inductors are mounted on the chassis in a conventional manner. An open-type power transformer and the autotransformer used for output voltage adjustment are mounted under

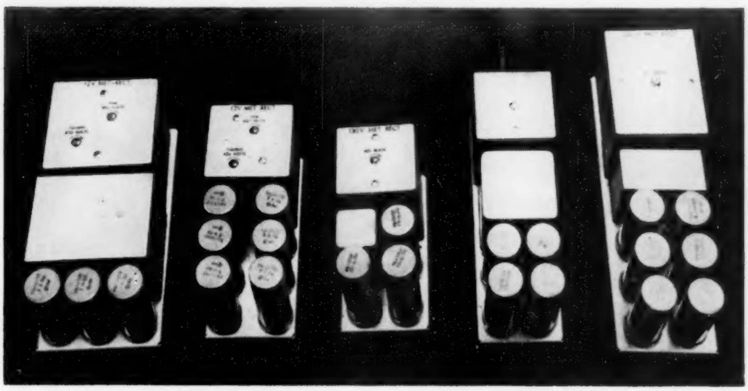

Fig. 5 — Rectifier packages showing family resemblance of designs.

a removable cover which is provided for personnel protection. The underside of the chassis is closed by a hinged cover which also mounts the silicon diode rectifiers. With the hinged cover closed all wiring except the input and output leads is protected. These leads may exit from the rectifier in either of two ways: from the side of the chassis or through the hinged bottom cover, depending upon the specific application of the rectifier. A typical rectifier is shown in Fig. 6.

4.2 *High-Voltage Traveling-Wave Tube Supply*

A supply with several special features was developed to power either one or two traveling-wave tubes. A summary of the high-voltage requirements is given in Table III.

A block diagram of the basic power supply is given in Fig. 7. The 1200-volt output is developed by a conventional unregulated rectifier and filter circuit. The independently adjustable helix and anode poten-

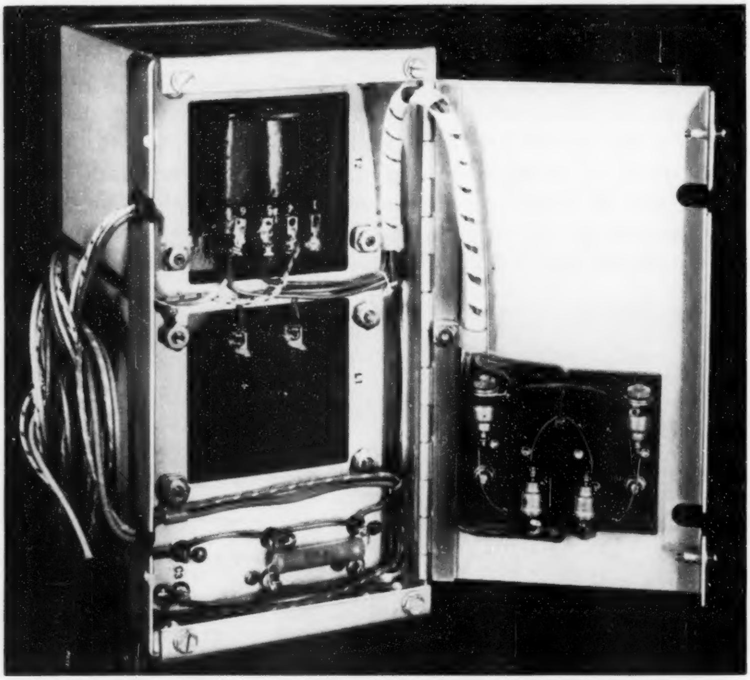

Fig. 6 — Typical unregulated rectifier, showing treatment of leads under hinged bottom cover.

TABLE III — HIGH-VOLTAGE REQUIREMENTS

Electrode	Voltage	Voltage Stability	Current Drain	Ripple (rms)
Anode	Adjustable, 2530 to 3100 volts	$\pm\frac{1}{2}$ per cent	0–1 ma	1.5 v
Helix	Adjustable, 2160 to 2610 volts or 1900 to 2200 volts	$\pm\frac{1}{2}$ per cent	0–1 ma	1.5 v
Collector	Fixed 1200 volts	± 2 per cent	45 ma	37.5 v

tials are derived by potentiometers connected to a regulated 2900-volt* dc bus. Feedback regulation is accomplished using a cold cathode gas tube for voltage reference, a two-stage electron tube amplifier, and a magnetic amplifier for control of the ac input voltage to a high voltage rectifier.

Silicon diffused junction diodes are used in a single-phase bridge circuit for high voltage rectification. Each arm of the 2900-volt rectifier bridge contains four diodes in series. The 1200-volt bridge has two series diodes per arm. Each diode consists of several diffused junction wafers bonded together by a thermal-compression technique. In this manner 2000 volts minimum reverse voltage blocking is achieved in a single

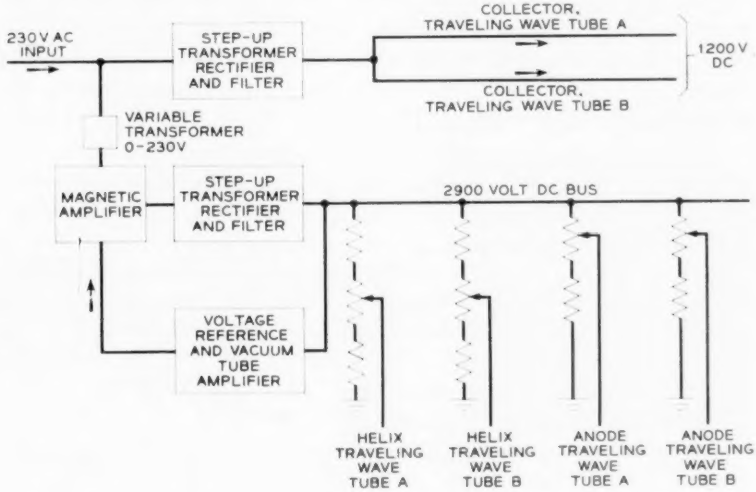

Fig. 7 — Block diagram of basic high-voltage power supply for traveling-wave tubes.

* This is the normal value. A few traveling-wave tubes require as high as 3100 volts on the anode.

two-terminal device. The transient peak inverse voltages due to random on-off switching of the ac input were carefully investigated. Peak transients of 3600 volts were found in the 1200-volt rectifier when ac power was removed at no load; the normal recurrent voltage peak is 2200 volts. Rather than add additional series diodes to block the 3600-volt transient peak, an RC network is connected across the transformer secondary winding. This eliminates transients in excess of the normal recurrent peak voltage.

After initial adjustment of the traveling-wave tube, the currents taken by the anode and helix electrodes are less than 1 ma, typically 0.1 ma. If, however, during the initial adjustment process or subsequent use the tube becomes defocused, destructively large current can flow to these electrodes. In addition, the major power dissipating element, the collector, which normally operates at 40 ma, can be damaged if the current exceeds 53 ma for a prolonged interval. To protect the tubes, an overcurrent detection and shutdown feature is provided on each of the six high-voltage outputs shown in Fig. 7. The circuit approach is outlined in Fig. 8. The current, I_H, delivered to the helix flows through the control winding of a small linear saturable reactor. The rectified output current, KI_H, of the saturable reactor is isolated from high voltage and is closely proportional to the helix current. This furnishes a safe and convenient means for measuring helix current at jacks J1 and J2. The current, KI_H, is the input signal to a bistable magnetic amplifier. The magnetic amplifier is normally biased such that no output signal exists

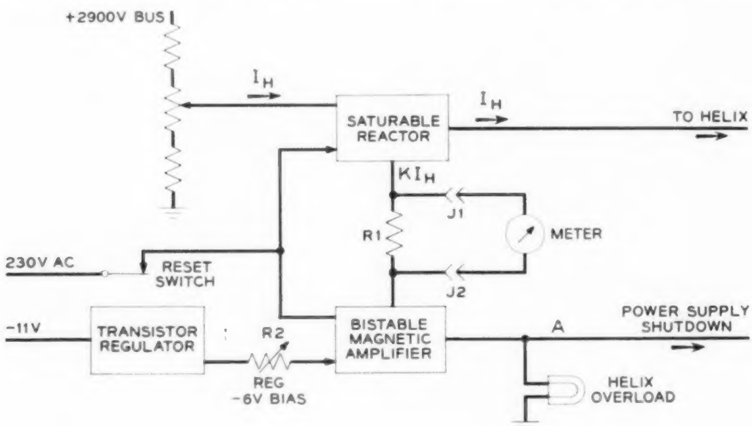

Fig. 8 — Circuit outline of overcurrent detection and shutdown feature.

on lead "A" for any helix current less than 2.5 ma. If, however, the helix current exceeds 2.5 ma, the bistable device is triggered on and a signal appears on lead "A" which automatically turns off the ac input power to the high voltage rectifiers. The helix overload signal remains on until the reset switch is actuated. The trip point is controlled by adjusting the bias current. The bias is derived from the 11-volt supply to the traveling-wave tube heaters and is stabilized at 6 volts by a transistor regulator. Should heater power be accidentally turned off, the loss of bias triggers on the bistable magnetic amplifier and protects the tubes by shutting down power. Similar circuits protect the anode and collector.

For further protection of the traveling-wave tube, a definite sequence is followed in the application of dc power. First the heater is energized, then the collector, and finally, after a five-minute warm-up interval, the helix and anode. The reverse order is followed for a normal turnoff. To facilitate initial tube alignment, a variable transformer ahead of the magnetic amplifier in Fig. 7 provides continuous manual control of the 2900-volt bus from zero to full voltage.

A three-position ac input rotary switch and a motor-driven timer are used to obtain the sequencing of the high voltages. No output is available in the first or OFF position of the switch. The second position turns on the 1200-volt collector supply. The third position starts a motor-driven timer which connects ac power to the 2900-volt rectifier after a five-minute delay. To avoid repetition of this delay on a momentary interruption of ac power, a 15-second pneumatic timer is arranged to bypass the five-minute delay if power is restored within 15 seconds.

The equipment design of the high-voltage power supply, shown in Fig. 9, presents a problem of packaging a mass of components operating at high potential within a small volume. This requires the recognition of areas where difficulties could arise due to dielectric breakdown, corona generation and personnel safety hazards. Also required is accessability to components to simplify maintenance and reduce the down time of equipment.

The entire supply is in a steel cabinet arranged for rack mounting. To utilize the volume available to the maximum extent, the dc supply is arranged in three packages consisting of a 1200-volt rectifier, a 2900-volt rectifier and a control unit which houses the ac input controls and the metering and overload protection circuit elements. As seen in Fig. 10, the individual equipment units are installed in the power supply cabinet using metallic slides on the chassis which mate with metallic guides in the cabinet. By disconnecting leads, units are removed individually for maintenance.

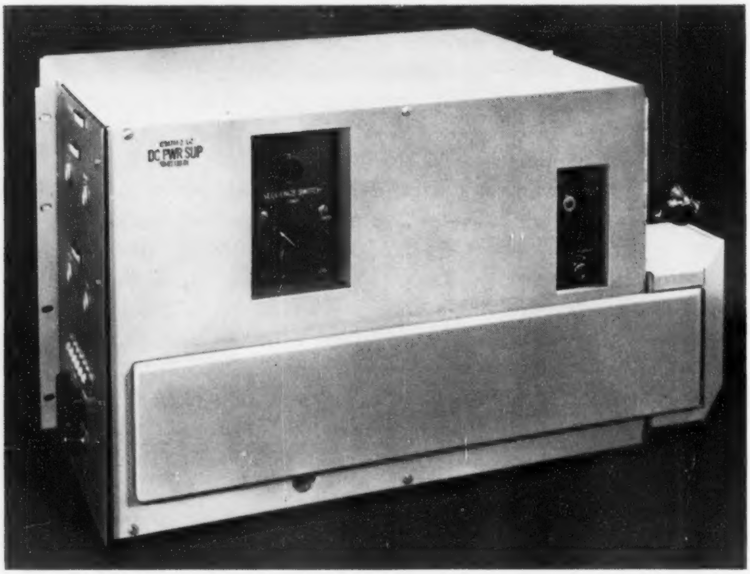

Fig. 9 — Equipment design of the high-voltage power supply.

To permit a compact mechanical design consistent with the high ac and dc potentials present, extensive use has been made of glass-fabric reinforced plastic molded chassis. This probably marks the first use of this construction for equipment chassis, although glass-mat covers have been used for some time on telephone answering sets and small power plants for subscribers' premises. These chassis make an important contribution toward the achievement of corona-free operation in equipments on a production basis. Forced air cooling is provided by a fan mounted on the side to minimize the temperature rise and obtain long life from the circuit components. In Fig. 10 the various equipment units may be seen partially removed from the cabinet. The control unit is at the left, the 2900-volt rectifier at the right, and the 1200-volt rectifier in the center.

In keeping with Bell System practices for personnel safety, several precautions have been observed in the equipment design of this power supply. The steel cabinet is arranged so that entry is not possible with the 1200 and 2900 voltages present. A mechanical key interlock system, similar to those used on TD-2 250-volt battery cabinets and the high voltage power supplies for the SB submarine cable systems, is employed.

The system requires that the rotary ac input switch previously described be locked in the OFF position before the door to the power supply or the cover to either of the two associated traveling-wave tube housings can be opened. Of course it is necessary that the door or covers be closed before ac power may be reapplied.

4.3 *Supplies for the FM Terminal*

Two power supplies were developed to meet the requirements of the FM terminal equipment. One of these supplies the heater, repellers and resonators of the two klystrons in the FM transmitter. The specific requirements are listed in Table IV. The heater requirement is satisfied with an adjustable, unregulated rectifier of the type described earlier. An electron tube regulated rectifier is employed to obtain +450 volts with high stability and low noise. The circuit consists of series triodes driven by a two-stage differential amplifier (electron tube) with cold cathode gas tubes for voltage reference. Until quite recently, it would not have been possible to achieve these operating requirements with a circuit of this simplicity because of the fluctuation and drift performance

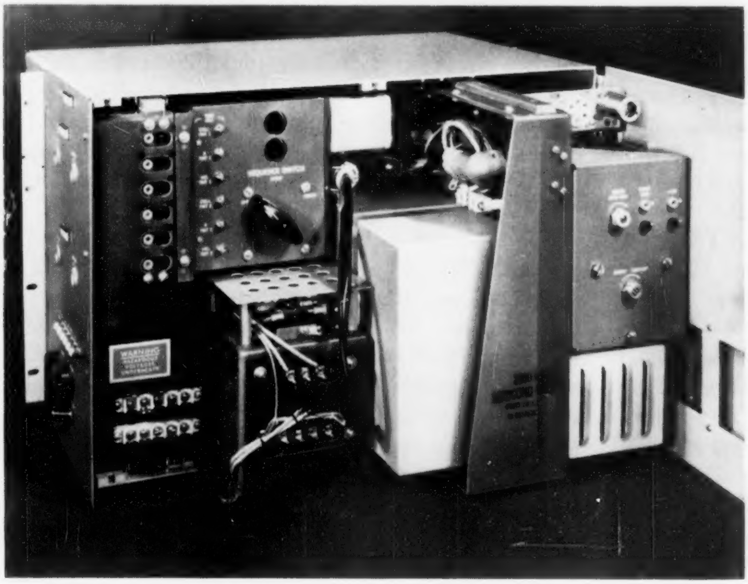

Fig. 10 — High-voltage power supply with various equipment units partially removed.

TABLE IV — REQUIREMENTS OF KLYSTRON SUPPLY

Tube Element	Voltage	Current	Voltage Stability	RMS Noise	
				$\frac{1}{4}$–30 cps	>30 cps
Resonator	+450	0.120 amp	±0.2 per cent*	0.5 mv	2 mv
Repeller	−170	0.004	±0.25 per cent*	0.7 mv	25 mv
Heater	−11	2.0	±2 per cent	17 mv	70 mv

* Limits apply to maximum voltage change due to line, load and 1000-hour drift.

of older voltage regulator tubes. However, tubes with adequate performance are now available.

The repeller output actually feeds two potentiometers in the FM transmitter from which independently adjustable repeller voltages are derived. It was not necessary, then, for the power supply to be adjustable to exactly −170 volts. By virtue of this factor and the constant nature of the load, it is possible to use a simple two-stage tandem VR tube regulator to satisfy the voltage stability and noise requirements.

The other FM terminal power supply furnishes the video and IF amplifiers, limiter, discriminator, AFC, etc. in either the transmitter or receiver. The similarity of the receiver and transmitter requirements led to the development of a power supply suitable for either application. The power performance is outlined in Table V. Both the −11-volt and +135-volt outputs are furnished by adjustable, unregulated rectifiers. An equally simple circuit could have been used for the +220-volt output except for the low noise specification. Instead, a conventional series tube regulated rectifier circuit was adopted.

The klystron supply is shown in Fig. 11. It is arranged for mounting on 19-inch cable duct frameworks. Opening the front doors permits access to the electron tubes and other components for maintenance purposes. Forced air from the bay cooling system is directed into the cabinet to supplement the cooling effect afforded by the grillwork in the supply doors. Within the power supplies extensive use is made of plastic grids for personnel protection.

TABLE V — PERFORMANCE OF TRANSMITTER-RECEIVER SUPPLY

Voltage	Current	Stability	RMS Noise	
			$\frac{1}{4}$–30 cps	>30 cps
+220	0.500 amp	±2 per cent	0.4 mv	2 mv
+135	0.375	±2 per cent	25 mv	15 mv
−11	10.0	±2 per cent	—	500 mv

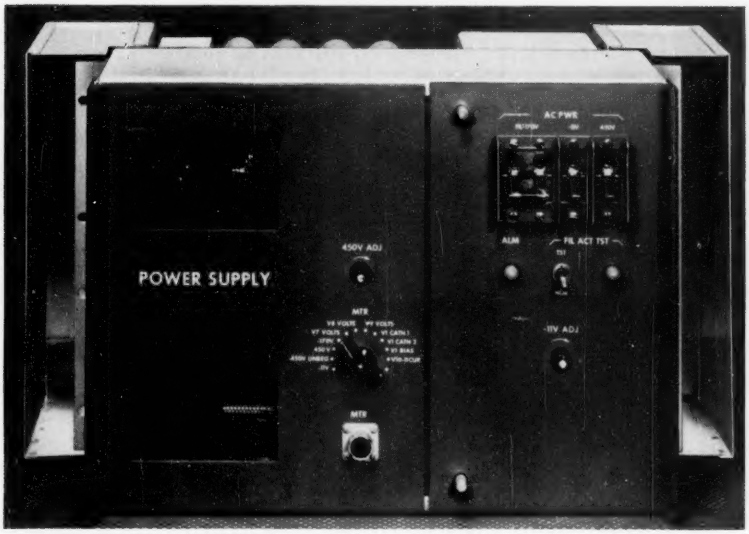

Fig. 11 — The klystron power supply.

4.4 *Power Plants for the Protection Switching Equipment*

A new power plant was developed to provide regulated +24 volts and −24 volts to the protection switching equipment. The requirements include:

Nominal voltage under full load	±23.9 volts
Long term voltage stability	±1.0 volt
Voltage deviation under emergency conditions	+1 volt to −3 volts
Ripple	<50 mv rms
Capacity	0–30 amp

This plant is designed on a centralized scheme using rectifiers on a building block basis for plant capacities of 10, 20 or 30 amperes for either the +24-volt or −24-volt applications. Magnetically regulated rectifiers are used with additional supplementary filtering to reduce the ripple level to within the specified limit. For an initial 10-amp load, two 10-amp rectifiers are furnished with each additional 10 amperes of load requiring an additional rectifier up to a maximum of four rectifiers per polarity. Each rectifier is connected to a separate firm ac bus for added reliability.

A relatively low capacity battery (100 ampere-hours) is provided to assure that the output voltage limits will be maintained under any tran-

sient or emergency condition, such as switching rectifiers on and off, failure of a rectifier, and transients on the firm ac bus.

The equipment for this plant is housed in self-supported eight-foot high floor mounted cabinets. The initial bay for each polarity contains the batteries, two rectifiers, controls, and charge and discharge fuses, and is rated at 10 amperes. The supplementary bay provides space for four additional rectifiers, two of each polarity, to build each plant out to its 30-amp capacity.

V. AUXILIARY ARRANGEMENTS

5.1 *Test Set Power Supplies*

In addition to the power arrangements described above, some special power supplies were developed for test equipment: (1) radio repeater test bench, (2) FM terminal test set and (3) protection switching test set.

The test bench is intended for use at maintenance centers remote from a working station. Commercial ac power is stabilized by a line regulator built into the test bench. With a regulated source of ac available, the dc power requirements are met by using dc power supplies already described for the radio equipment. The line regulator is rated at 5 KVA and arranged for nominal 208- or 230-volt input and 230-volt output. Regulation is by a motor-driven, variable transformer which is coupled to a "buck-boost" transformer in series with the output.

The FM terminal test set is mounted in a mobile bay and operates from convenience outlets for 117-volt ac power. Plate voltage is provided by two series tube regulated rectifiers which are adjustable from 200 to 300 volts dc and rated at 0.1 ampere. A ferroresonant regulated power supply furnishes heater power at 6.4 volts ac, 15 amperes, and 6.4 volts dc, 1.5 amperes. At constant load, the heater voltage is constant within ± 1 per cent for a ± 10 per cent variation in input voltage.

The protection switching test set, also mounted in a mobile bay, is equipped with two regulated rectifiers which furnish $+24$ and -24 volts. A single design suitable for either the positive or negative application is used. The output voltage is adjustable from 23.5 to 24.5 volts dc; it is stabilized within ± 1 per cent for ac input voltages of 117 volts ± 10 per cent, frequency 60 cps ± 2 per cent, and load 0 to 1.5 amperes. The output ripple is 40 mv rms. Regulation is accomplished with a series transistor, driven by a two-stage transistor amplifier and a silicon voltage reference diode. A ferroresonant regulator is built into the stepdown transformer preceding the rectifier portion of the circuit. This stabilizes

the dc input voltage to the regulator against line voltage changes and reduces the maximum series transistor dissipation from 12.5 to 4.5 watts. The rapid decrease in output voltage with overload currents, characteristic of ferroresonant regulators, coupled with a bias circuit which maintains the series transistor in a saturated state, protects the series transistor from excessive dissipation if an overload is applied to the regulated dc output terminals.

5.2 *Lightning and Grounding Arrangements*

Radio and microwave relay facilities, by their nature and location, are vulnerable to lightning strokes to their antennas and towers, much more so than local and toll central offices. Therefore, at radio and microwave relay stations all the established engineering practices to achieve central grounding are supplemented by extensive measures to provide lightning protection for the operators of the equipment, the equipment and the building.

The effectiveness of the protection afforded to equipment and personnel is primarily a matter of creating a unipotential area by extensive interconnection of all components within the area. A basic grounding arrangement is shown in Fig. 12. At repeater stations the practice has

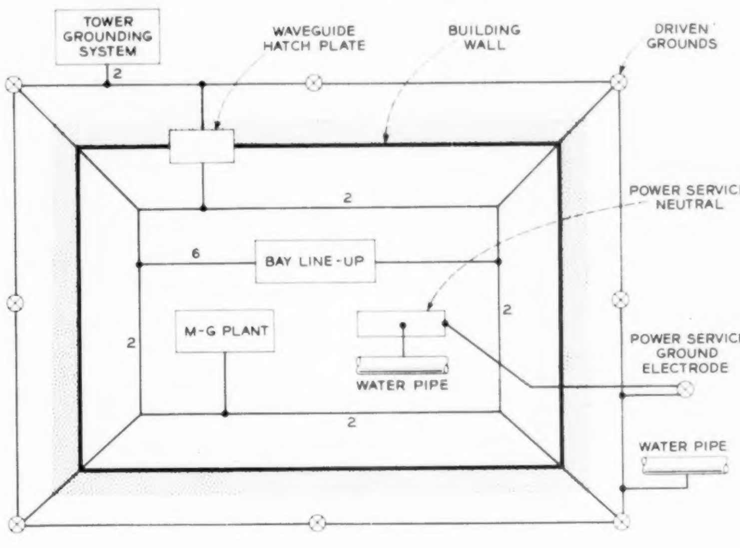

Fig. 12 — Arrangement of the basic building grounding system.

been adopted of providing a buried ring around the exterior of the building using #2 bare copper wire with driven grounds at intervals of ten feet. A similar ring within the building is provided at a convenient height, mounted on the walls or supported from cable racks; it is connected to the outside buried ground ring at several points. The ends of various equipment lineups are tied to this internal ring with at least #6 wire. These include the engine control bay, 508A motor alternator plant, 130-volt power bays, TH Radio Bays, tube cooling system equipment, conduits, miscellaneous radio bays and waveguides. The grounding arrangements for the self-supported tower at repeater stations are tied into the buried ground ring outside the repeater building to direct lightning strokes to ground by the shortest path. Unusual measures are taken to ground the waveguides at the waveguide hatch plates to both the internal and external ground rings. When a water pipe system is present it is bonded to the buried ground ring.

Power service to TH stations is of the three-phase four-wire type because of the heavy load. Two stages of secondary lightning protection are provided for the service leads. Secondary arrestors rated at 1750 volts to ground are connected from each phase to ground at the weatherhead on the service entrance. These arrestors have a spark-over value of 1600 volts. Branch circuit arrestors are connected across branch secondary circuits on the load side of low-capacity fuses. These branch circuit arrestors consist of nonlinear elements in series with a gap which sparks over at about 850–1000 volts. To insure the operation of the 1600-volt arrestors, the service leads must be in at least twenty feet of grounded conduit between the 1600-volt arrestor and the branch circuit arrestors. This length of conduit at lightning frequencies provides a beneficial choking effect to surge currents.

The power service neutral is grounded to the buried ground ring or water pipe system, when available, or to a separate grounding electrode if required by the power company. These three, if all are present, are bonded together.

Auxiliary Radio Channels for the TH Radio Relay System

By R. W. HATCH and P. R. WICKLIFFE

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The operation and maintenance of the TH and TD-2 radio relay systems require communication facilities along the route which can be used for the transmission of voice, protection switching and alarm signals. Special narrow-band microwave facilities, referred to as the auxiliary radio channels, which have been developed as an integral part of the TH system to supply this need, are described in this paper.

I. INTRODUCTION

The operation and maintenance of the TH and TD-2 radio systems require communication facilities along the route which can be used for the transmission of voice, protection switching and alarm signals. In the TD-2 radio system such facilities are typically provided by wire lines running approximately parallel to the radio route, but often separated by several miles. Where such separations exist, additional wire-line facilities are required to connect the individual radio stations with the parallel wire-line route. The desirability of providing narrow-band microwave channels for the transmission of these signals within the frequency allocation of the TH system was recognized early in the TH development. Narrow-band channels which have been designed for this purpose are described in this paper. These channels, called auxiliary channels, serve both the TD-2 and the TH systems where there is a combined installation.

Before proceeding with a detailed discussion of the auxiliary channel system, its associated systems and signals will be briefly described.

1.1 *The C1 Alarm and Control System*

The C1 alarm and control system as it was originally developed for use with the TD-2 system has been described in a paper on that system.[1]

Its purpose is to allow an attendant at an alarm center to monitor the operation of a number of unattended radio stations. An expanded but otherwise similar version now exists which can handle a combined installation of TD-2 and TH. A maximum of twelve unattended radio stations can be associated with a single alarm center. These are usually arranged so that not more than six stations are located on either side of the alarm center. A one-way voice-frequency circuit (known as the alarm line) is provided from each group of six stations to the alarm center. Over this circuit each station transmits a continuous and distinctive tone at one of the six frequencies (1100, 1300, 1500, 1700, 1900, and 2100 cycles) which are available for this use. These tones are detected in separate frequency selective circuits at the alarm center. The interruption of a tone for approximately ten seconds registers an audible and visual alarm. Because each station is assigned a different frequency, the station whose tone is interrupted is easily identified.

The removal of one of the station alarm tones described above indicates that a particular unattended station is in trouble, but it does not identify its nature. To get this detailed information, the attendant at the alarm center sends an order over another one-way voice circuit* (known as the order circuit) which connects to the group of six stations. The orders consist of a 1600-cycle carrier modulated in sequence by selected combinations of twelve modulating frequencies available in 16-cycle steps from 277.5 to 442.5 cycles. By choosing a particular combination, one of several possible orders can be sent to a particular station. Typical orders might be to start the gas engine alternator or, as in this case, to scan certain alarm indications. As these alarm indications are scanned in the station in response to an order, a series of 900-cycle pulses are transmitted at a 5-cycle rate over the two-wire alarm line back to the alarm center. Whenever an alarm condition is encountered, a 700-cycle pulse is transmitted simultaneously with the 900-cycle pulse. The presence of a 700-cycle pulse causes the appropriate lamp to be lighted on a lamp display in the alarm center.

1.2 *Voice Circuits*

A second two-way, four-wire voice circuit is used for telephone conversations which are required in connection with system maintenance. It connects all the stations along a section of the radio route with the associated alarm center. Known as the "radio order circuit," this circuit is essentially a party line which permits an individual in one station

* This one-way circuit is opposite in direction to the alarm circuit. Thus, the other half of a four-wire, two-way telephone channel can be used for this purpose.

to talk with an individual at any of the other stations in the section. Means for signaling from one station to another are provided as a part of the C1 system.

Additional voice circuits, called "express radio order circuits," are provided over much longer sections of the radio route, but only selected stations are connected. Normally, these "express circuits" will not be provided over the auxiliary channels.

1.3 *TD-2 Protection Switching*

The TD-2 protection switching system has been described in an earlier paper.[2] Two one-way voice channels are required; each permits transmission from the receiving end to the transmitting end of a TD-2 switching section, a distance which normally consists of from five to fifteen radio hops. When all of the regular radio channels in the section are working properly, a 700-cycle guard tone is transmitted over this facility. This guard tone prevents the operation of TD-2 protection switches on extraneous signals, such as noise in the voice channel, and provides an indication that the voice-frequency facility has not been impaired. If detectors at the receiving end indicate that transmission on one of the regular TD-2 radio channels is impaired, and if the protection channel is available, the 700-cycle guard tone is removed and another tone, depending on the channel in trouble, is transmitted. Tones at 900, 1100, 1300, 1500 and 1700 cycles are used for the five regular channels. When the tone corresponding to a particular channel is received at the transmitting end of the section, the signal on the regular channel is transferred to the spare channel. When transmission on the regular channel returns to normal, the procedure is reversed: the tone representing the regular channel is removed and the 700-cycle guard tone is restored.

1.4 *TH Protection Switching*

The TH protection switching system described in a companion paper[3] requires transmission facilities for a total of sixteen tones in each direction along the radio route. These tones are located at 1-kc intervals in the band from 20.5 kc to 35.5 kc. Of the sixteen tones transmitted in one direction, eight alternate tones starting at 21.5 kc are status tones for the eight broadband channels being transmitted in the opposite direction. The other eight (actually two groups of four) are order tones for the eight channels being transmitted in the same direction as the tones. A two-out-of-four code is used for each group of order tones.

Thus, for the eight channels in a given direction there are eight possible order tones in the same direction and eight status tones in the opposite direction.

Tone-reporter circuits are provided in each radio station which can selectively attenuate any of the eight status tones. If the transmission on a particular broadband channel is impaired, the corresponding status tone is attenuated. This action is detected at the transmitting end of the protection switching section, and a switching action will be initiated there. The order tones, in the other direction, are then used to signal for the corresponding switch at the receiving end.

II. DESIGN FEATURES

To compete with wire-line facilities, the auxiliary channels must provide a less expensive and more reliable medium for transmitting the signals described in the preceding sections. As a result, the auxiliary channels share the antenna system and common microwave carrier supply of the broadband channels. In addition, the auxiliary channels employ double-sideband amplitude modulation, operate without radio frequency amplification in the transmitter, and use components designed for the broadband channels wherever practical.

The required reliability is achieved by providing two narrow-band radio channels for each direction of transmission. At each radio station, the baseband inputs and outputs of the two channels are connected in parallel. In addition, the automatic gain control circuits of the receivers operating in the same direction are connected in parallel. This interconnection of baseband circuits and automatic gain control circuits provides protection against fades in the transmission path and equipment failures without the necessity of automatic switching circuits and results in a nearly optimum signal-to-noise ratio. This equipment diversity has the further advantage of permitting in-service removal of units for maintenance.

At each station, it is necessary to recover the baseband signal so that specific components, as described in the preceding sections, can be added or removed. For a ten-hop route — the maximum length of a TH protection switching section — a total of ten amplitude modulators and demodulators are thus connected in tandem. To reduce the linearity required in these circuits, a baseband frequency allocation is used in which second-order modulation products fall outside the signal bands. This leads to the baseband signal spectrum shown in Fig. 1. Some further advantage in over-all linearity is achieved by modulating the two channels in a given direction with baseband signals which are 180° out

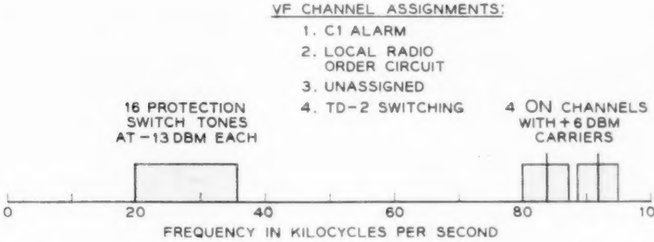

Fig. 1 — Baseband signal spectrum for auxiliary radio channels.

of phase so that even-order modulation products tend to cancel when the baseband signals are recombined at each station.

III. DESCRIPTION

3.1 *General*

A simplified block diagram of a typical auxiliary channel installation is shown in Fig. 2. In the west-east (W-E) circuit shown at the top, amplitude modulated microwave signals are received on channels 10 and 19.* Radio receivers with interconnected AGC circuits (which will be discussed in greater detail in Section 3.3) recover the baseband signals, which are then combined in a hybrid. This combined baseband signal is then fed to a transmitting hybrid where the signal is again split and applied in parallel to the two radio transmitters on channels 20 and 29. The receiving portion of the ON carrier terminal is bridged across the baseband circuit. This permits any of the four telephone signals on the W-E circuit to be received at the respective voice-frequency terminals of the ON equipment. In addition, a tone-reporter circuit (Section 1.4 above) is bridged across the baseband connection between receivers and transmitters. Signals from the transmitter portion of the other ON carrier terminal are added to the baseband signal by means of the connection shown to the right of the tone reporter. The lower half of Fig. 2 shows a duplication of these facilities for the other direction of transmission.

In the arrangement described above, the baseband signals, including the four ON channels in the 80–96 kc band, are connected directly from the radio receivers to the radio transmitters. In this way signals being

* The frequency allocation of the auxiliary channels is described in detail in Ref. 4.

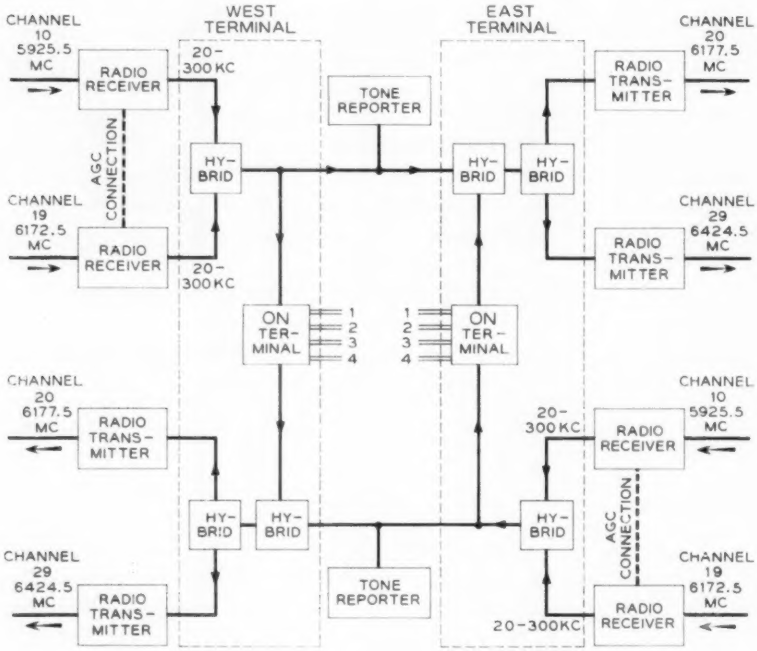

Fig. 2 — Simplified block diagram of typical auxiliary channel repeater installation.

transmitted through a given station are not impaired by repeated demodulation to voice frequency and by the subsequent modulation back to baseband frequency. However, the ON terminals are connected in such a way that signals can be received and transmitted on any of the voice channels in either direction. This arrangement provides certain advantages. Transmission impairments for ten ON terminal pairs in tandem are avoided. This is particularly desirable for the compandored circuits. Furthermore, only those voice circuits required at a given station need be equipped. On the other hand, this method of operation requires the twin channel carriers at 84 kc and 92 kc in the ON terminal to be suppressed at all except the first station in a section. At all other stations, twin-channel sidebands are added in the baseband signal adjacent to the twin-channel carriers which originate in the first station. Therefore voice-frequency signals originating at other than the first station are subject to a slight frequency shift equal to the difference between the twin-channel carrier at the first station and that of the station where the signal originates.

3.2 *Radio Transmitter*

A block diagram of the radio transmitter is shown in Fig. 3. The baseband signal is combined with a microwave carrier in the first modulator — the transmitter modulator — to produce a double-sideband, amplitude modulated signal. This microwave signal, in turn, is shifted to the assigned channel by combining it with the output of a crystal-controlled oscillator in a second modulator. The output of this second modulator — the frequency shift modulator — contains amplitude modulated carriers corresponding to both the sum and the difference of the two input frequencies. The bandpass filter selects the desired signal. The isolator following the bandpass filter provides good terminations for both the filter and the channel-separation network through which the transmitter is connected to the antenna system.

3.2.1 *Transmitter Modulator*

The transmitter modulator, shown in Fig. 4, consists of a dual isolator, a short-slot directional coupler, a diode mount, and a baseband coupling network. It is closely related to the broadband transmitter modulator, described in a previous article,[5] with adaptations for amplitude modulation use. The waveguide attenuator and the cross-guide directional coupler are used to measure and adjust the microwave input power. Fig. 5 is a simplified schematic of the transmitter modulator.

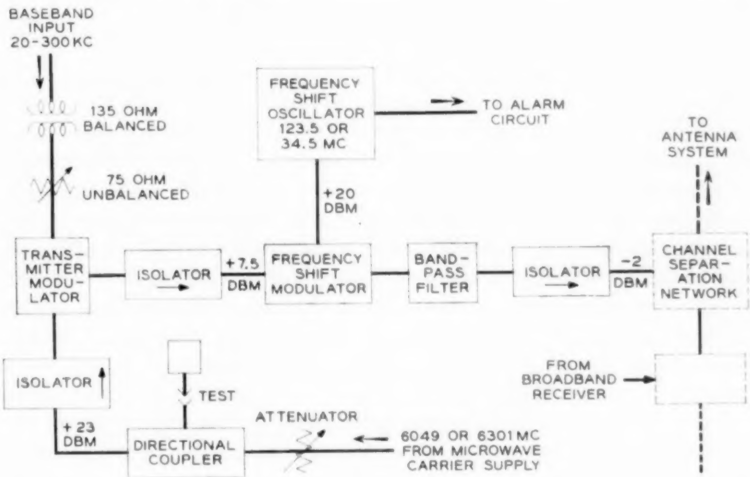

Fig. 3 — Block diagram of auxiliary channel transmitter.

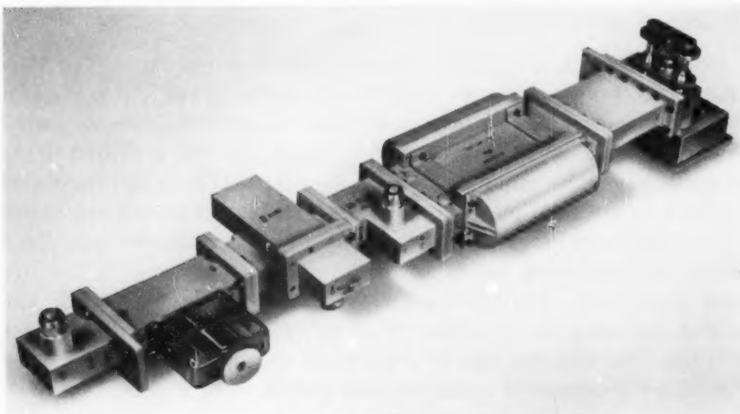

Fig. 4 — Modulator of auxiliary channel transmitter.

The short-slot directional coupler divides the incoming microwave carrier equally between the two diodes CR1 and CR2. Each diode absorbs a fraction of the incident carrier and reflects the remainder. The directional coupler divides each reflected signal equally between the IN and OUT ports. However, the phase characteristics of the directional coupler cause these reflected signals to add at the OUT port and to cancel at the IN port. As a result, the signal at the OUT port is proportional to the sum of the complex reflection coefficients of the diodes, and the signal at the IN port is proportional to their difference. The reflection coefficients are determined by the microwave tuning adjustments (RF TUNERS 1 and 2) and the diode biases. When these tuning adjustments are properly set, the magnitude of the reflection coefficient varies with the diode bias, and its angle remains constant. For a given carrier power there is a value of diode bias about which the magnitude of the reflection coefficient varies linearly with the diode bias. If the low-frequency baseband signals are superimposed on this dc bias, the reflection coefficients and, in turn, the signal at the OUT port are amplitude modulated. The dc bias determines the carrier power and the amplitude of the superimposed baseband signals determines the per cent modulation. Variations of the microwave carrier input produce variations in both the microwave carrier output and the per cent modulation. To minimize these variations, the diodes are operated with a combination of self- and fixed-bias. Rectified currents flowing through R1A and R1B produce the self-bias; the fixed-bias is developed across R2. Inductors L1 and L2

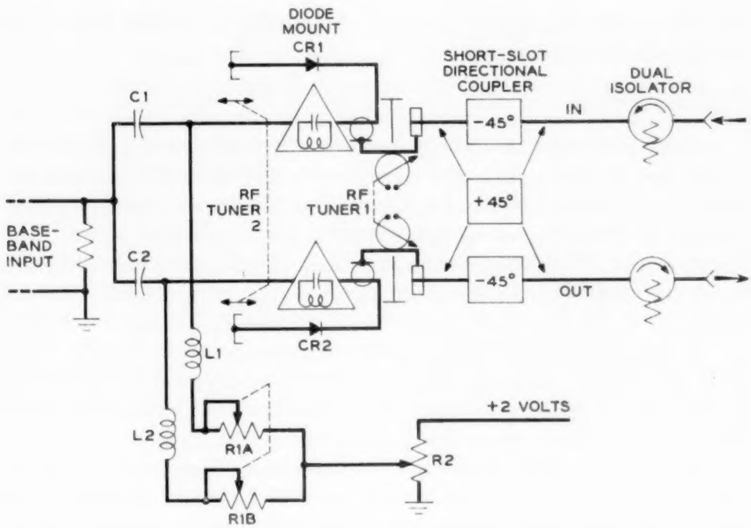

Fig. 5 — Simplified schematic of the transmitter modulator.

and capacitors C1 and C2 separate the baseband circuits and the bias circuits.

The frequency shift modulator is similar in appearance and operation to the carrier supply modulator described in a companion paper.[5] The powers shown in Fig. 3 indicate the performance of this modulator.

3.2.2 *Frequency Shift Oscillator*

The frequency shift oscillator is a conventional oscillator-buffer amplifier combination. The frequency determining element in the oscillator stage is a series-resonant overtone crystal. Normally, the frequency is 123.5 mc. However, when only one antenna is provided for each direction of transmission,[4] the channel 20 transmitter is temporarily assigned to channel 23, and the frequency of the associated oscillator is changed to 34.5 mc. Rectified current from a diode power monitor in the output circuit of the buffer amplifier operates a sensitive meter-type relay. A reduction in the output power of the buffer amplifier causes the relay to operate and, in turn, to initiate an alarm. Since the microwave carrier supply contains low-output alarms for the 6049-mc and 6301-mc carriers connected to the transmitter modulator, the inclusion of a similar alarm in the frequency shift oscillator insures, in effect, that any

significant reduction in the output of the auxiliary channel transmitter will initiate an alarm.

3.3 *Radio Receiver*

A block schematic of the radio receiver is shown in Fig. 6. The incoming microwave signal, which has been selected by the channel separation network, is connected to the receiver modulator through a bandpass filter. The receiver modulator shifts the microwave signal to an IF of 64.2 mc by combining the incoming signal with a beat-oscillator frequency obtained from the common microwave carrier supply system, exactly as in the broadband radio receiver.[5] The shift frequency of 59.3 mc is normally used with the carrier supply modulator; 29.65 mc is used when the channel 20 receiver is temporarily assigned to the channel 23 slot. This temporary assignment occurs at those stations equipped with only one antenna for each direction of transmission.

The 64.2-mc output of the receiver modulator is amplified in a three-stage preamplifier and is connected to the IF amplifier through an IF attenuator. This fixed pad is used to adjust the input power to the IF amplifier and to compensate for differing free-space path losses between repeater stations (e.g., short hops). The first three stages of the IF amplifier are controlled by a separate automatic gain control (AGC)

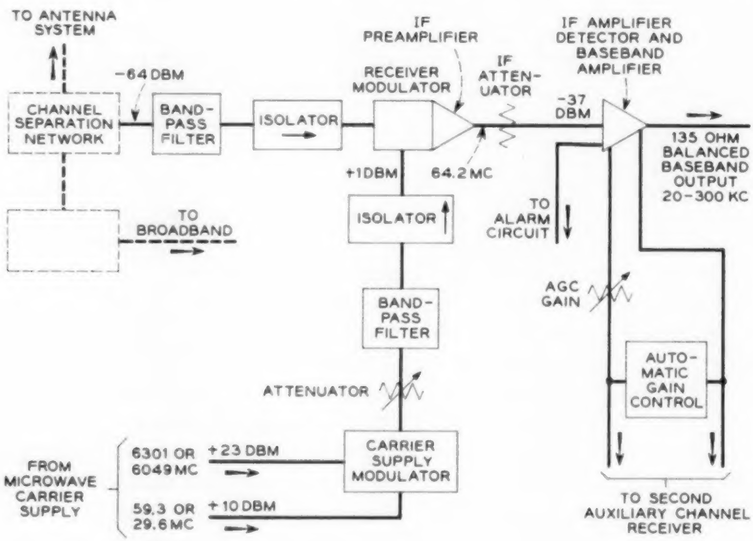

Fig. 6 — Block diagram of the auxiliary channel receiver.

unit; the fourth stage operates at fixed gain. The output of the fixed-gain stage is applied to an averaging detector which recovers the original baseband signal and also provides a voltage proportional to the applied IF signal. The baseband output of the detector is amplified in a two-stage feedback amplifier and then connected to the terminal equipment, where it is combined with the baseband output of the parallel radio receiver. The dc output of the detector is connected to the AGC circuits.

In normal operation (paralleled radio channels) the two AGC units are operated in parallel. Fig. 7 shows the interconnection. The internal impedance of each detector loads the other so that the actual voltage applied to the AGC units is one-half the sum of the open-circuit voltages of the two detectors. The outputs of the AGC amplifiers, Fig. 7, are also combined. The diodes permit the AGC unit with the more negative output voltage to control both IF amplifiers. The action of this interconnection is to maintain the sum of the receiver baseband outputs practically constant. The AGC GAIN adjustments compensate for differences between the individual loop gains.

The switches shown in Fig. 7 are arranged so that one receiver may

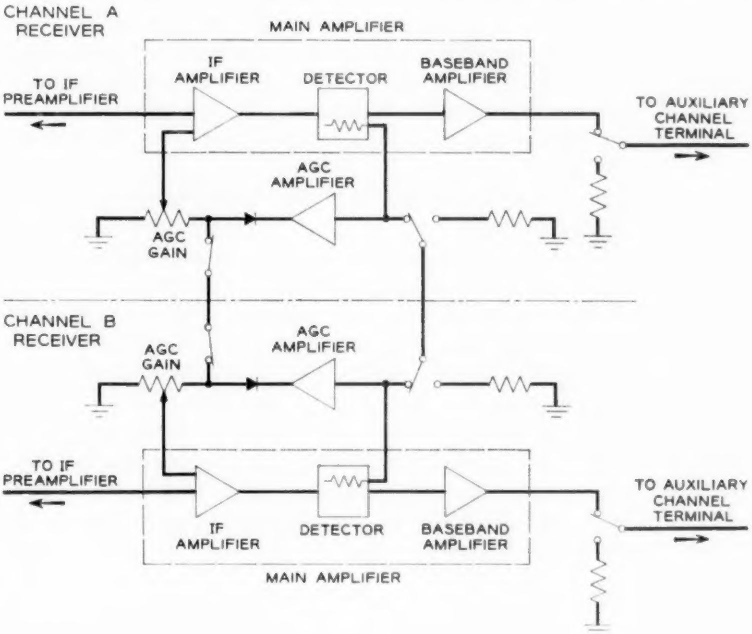

Fig. 7 — Block diagram showing interconnection of AGC circuits.

be disconnected without affecting the baseband signal level in the terminal equipment When the switches associated with one receiver are operated, the interconnections are broken, and the detector of the operating receiver is terminated. With only one detector contributing to the input of the remaining AGC unit, the gain of the corresponding IF amplifier increases 6 db to maintain that dc output constant. The baseband output of the operating receiver also increases 6 db and, as a result, the sum of the two receiver outputs remains constant.

The receiver modulator in the auxiliary channel receiver is identical to the modulator used in the broadband receiver, but the preamplifier is slightly different. The auxiliary channel preamplifier utilizes two 417A triodes in a modified cascode circuit followed by a 404A pentode as an output stage. The net gain of the receiver modulator-IF preamplifier is 37 db, and the overall noise figure is typically less than 10 db.

3.3.1 *IF Amplifier-Detector and Baseband Amplifier*

The three gain-controlled stages of the IF amplifier employ 404A pentodes and the fixed-gain stage, a 418A pentode. Each variable-gain stage has a cathode compensation network to stabilize the gain-frequency characteristic as the bias on the stage is varied. The detector employs two 427A gold-bonded germanium diodes connected in a voltage doubler configuration. The gain of the IF amplifier is sufficient to maintain a 2.5-volt dc detector output when the input signal is between -35 dbm and -60 dbm. The transmission characteristic is flat over a 2-mc band centered on 64.2 mc.

The two-stage baseband amplifier which follows the detector employs a 404A as a voltage amplifier and a triode-connected 418A as an output stage. With feedback, the transmission characteristic of the baseband amplifier is flat to ± 0.1 db from 20 kc to 200 kc. The transformer-coupled output stage is capable of delivering $+21$ dbm to a 135-ohm balanced load with negligible distortion.

As a means of indicating multipath fades and equipment failures, the 84-kc twin-channel carrier of the ON multiplex equipment is sampled by a narrow-band filter bridged across the output of the baseband amplifier. The output of the filter is amplified and detected, and the rectified current operates a sensitive meter-type relay. If, as a result of a fade or equipment failure, the 84-kc twin-channel carrier drops 20 db or more, an alarm is initiated.

3.4 *Terminal Circuits*

The terminal circuits at a typical repeater installation have been discussed in Section 3.1 with reference to Fig. 2. Although this represents

the most common case, other arrangements are required at the ends of a TH protection switching section and at intermediate points where one or more of the voice circuits are terminated. The latter may occur at alarm centers or the ends of TD-2 protection switching sections. These special interconnecting arrangements are shown in Fig. 8.

Standard ON terminals, described in Ref. 6, are used in all cases. In some respects, however, the method of operation differs from conventional ON practice. Most significant is the previously described suppression of the twin-channel carriers at intermediate stations. This is easily accomplished since balanced modulators are used in which the carriers are suppressed. The carriers are later inserted under control of a potentiometer, which can be set for zero carrier. Another change consists of disabling the signaling circuits, which are not now required because signaling is done over the C1 alarm system.

Several types of channel units are available for use in the ON terminals. Channel units with compandors are used for the voice circuit in order to improve the noise performance. Special service channel units are used on the circuits for C1 alarm and TD-2 protection switching.

At the ends of TH protection switching sections, the east terminal

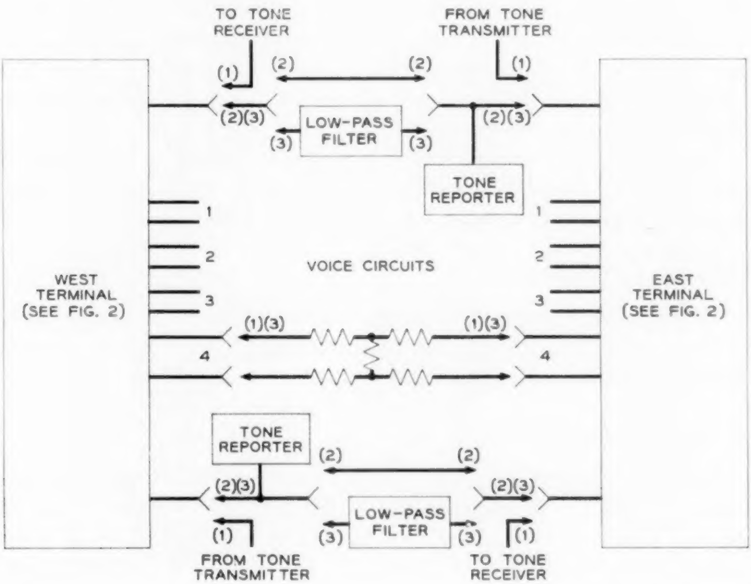

Fig. 8 — Simplified block diagram showing special terminal interconnections.

and the west terminal are separately connected to the appropriate tone transmitters and receivers in the TH protection switching system. In this arrangement, shown as option 1 in Fig. 8, there is no baseband connection between the two terminal circuits. The east and west sides of the stations are served by separate auxiliary channel systems. When it is necessary to provide continuity in a voice channel through such a station, the appropriate connections are made at voice frequency, as shown for voice circuit No. 4. In such stations the ON twin-channel carriers at 84 kc and 92 kc are supplied to the baseband circuit by the ON carrier terminals. At the end of a TH route a similar arrangement is used except that only one of the two terminal circuits needs to be provided.

At typical intermediate stations in a TH protection switching section, the arrangement shown as option 2 is used. This is identical with Fig. 2, already discussed. At such stations the ON twin-channel carriers are suppressed.

Still another arrangement is used, shown as option 3, at intermediate stations in a TH protection switching section where one or more of the voice circuits must be terminated. Such a situation can exist at the ends of C1 alarm sections or at the ends of TD-2 switching sections. In this case, a low-pass filter is provided in the baseband circuit which blocks the four ON channels but allows TH protection switching signals to pass through. The ON carriers must be provided at such stations, and all voice circuits which require continuity through the station require voice-frequency connections.

Fig. 9 shows a photograph of the auxiliary channel terminal bay. A photograph of the radio transmitter-receiver units and a more complete description of the equipment design are given in Ref. 7.

IV. PERFORMANCE

The performance data discussed in the following paragraphs are based on field measurements of an eight-hop loop and on laboratory measurements of several prototype units.

4.1 *Transmission Gain*

Transmission gain, measured from the common input of two paralleled radio transmitters to the combined output of the corresponding receivers, is essentially constant between 20 kc and 200 kc. Field measurements showed the following results: 20 kc, -0.2 db; 50 kc, 0.0 db; 200 kc, -0.3 db.

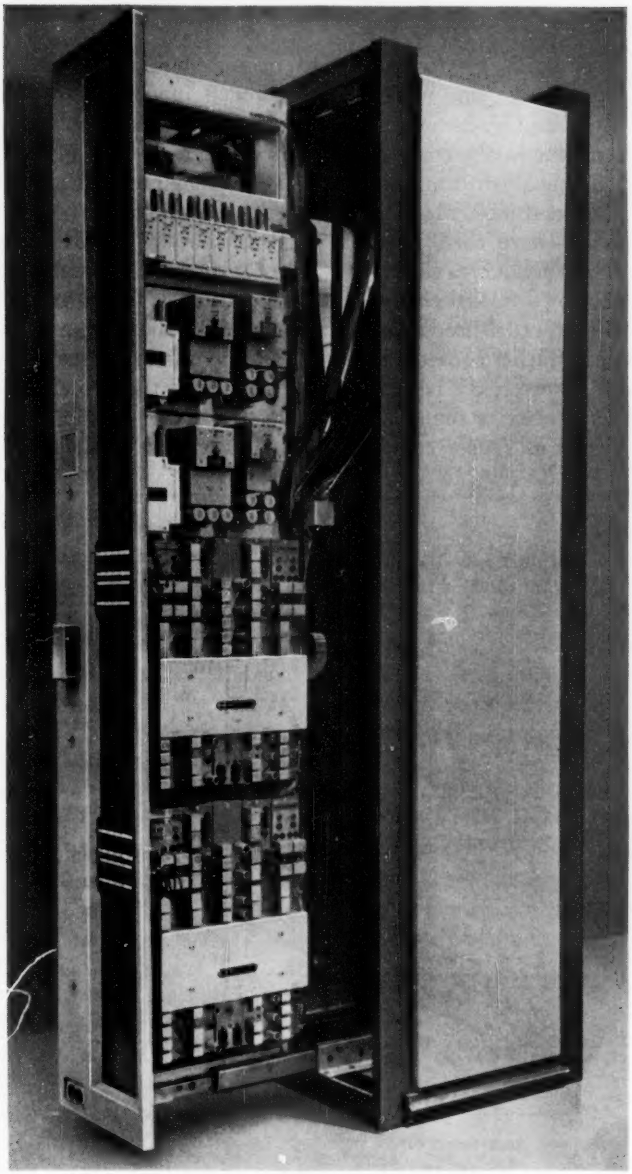

Fig. 9 — The auxiliary channel terminal bay.

4.2 *Modulation*

The modulation performance was measured in the laboratory by applying a 100-kc sine wave at the common input to two paralleled transmitters and by measuring the fundamental, second harmonic, and third harmonic at the combined output of the corresponding receivers. A plot of typical performance obtained in this manner is shown in Fig. 10. Of particular interest are the curves of second harmonic performance. The dashed curve corresponds to exciting the two radio channels in phase. The solid curves show the effect of reversing the baseband phase of one of the paralleled channels at both ends so as to get some cancellation of even-order products. As can be seen, the improvement is quite pronounced during nonfading periods. During a fade on one of the two channels, however, the baseband contributions of the two channels are unequal and some of the advantage is lost.

Field tests at 70 per cent modulation showed that the second har-

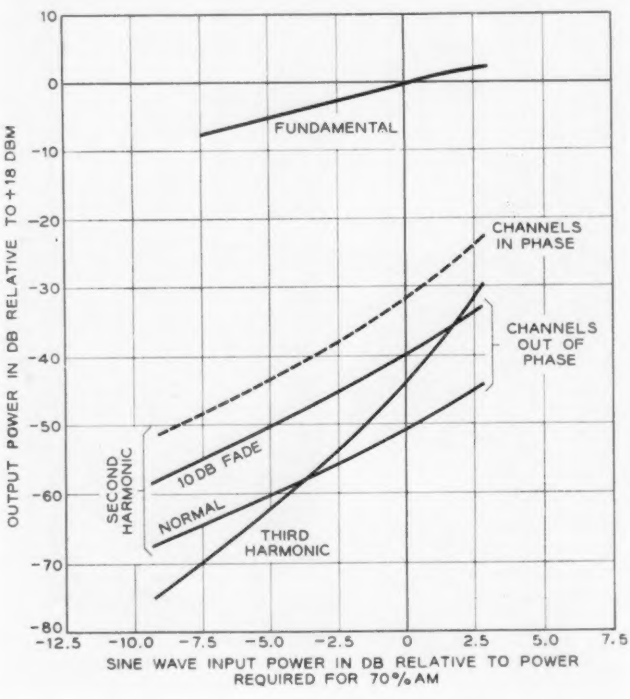

Fig. 10 — Typical modulation performance of a one-hop system.

monic power at the end of the eight-hop loop was 42 db below the fundamental and that the third harmonic was 34 db below the fundamental. The ratio of second harmonic to fundamental is well within system objectives; the objective for the ratio of third harmonic is 22 db.

4.3 *Fluctuation Noise*

Fluctuation noise in the baseband output depends on a number of factors, the most significant of which are transmitted power, receiver noise figure, and the radio path loss. The nominal transmitted power is −2 dbm, the receiver noise figure is 10 db, and the nominal loss from the transmitter output to the receiver input is 62 db. The fluctuation noise in a 3-kc band at combined output of the radio receivers is expected to be −44 dbm. Since this is a −2 db transmission level (TL) point, the noise would be −42 dbm or 40 dba at a 0-db TL point. At the end of 10 repeaters the noise would be 10 db greater or 50 dba. Thus, in the compandored voice circuits which have a subjective noise improvement of approximately 23 db, the apparent noise will be 27 dba at 0-db TL during nonfading periods.

The field data for the 8 hops gave noise power as −50 dbm in a 500-cps band at −8 db TL. This is −34 dbm in a 3-kc band at 0 db TL and agrees well with the expected value.

The bandwidth of the selective filters for the protection switching tones is approximately 400 cps. At 0-db TL the expected noise in this band is −41 dbm at the end of 10 repeaters. Since the tones have a power of −13 dbm at this point, the nominal signal-to-noise ratio for a protection switching tone is 28 db.

A statistical study based on fading and failure data which have been accumulated for the TD-2 system indicates that for the worst fading periods the performance will be somewhat poorer. However, for 90 per cent of the time during the midnight to 6 A.M. period during the worst three fading months, the expected performance will be degraded less than 4 db. For 99 per cent of the time during the same period the expected degradation is less than 7 db.

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The Automatic Protection Switching System of TH Radio

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It is the purpose of the automatic protection switching system to ensure continuity of service over the TH system. The protection system prevents any signal loss from occurring in case of fading, and it will restore service a short time after transmission equipment has failed. The over-all reliability of the TH system is thereby very substantially improved. Switching is done in switching sections. These usually contain a number of repeater stations, in which case the switching is normally done at IF. When a channel has to be switched, it is replaced over the full length of the switching section by a protection channel. Protection is also provided for FM terminal transmitters and receivers; in this case the switch is at IF on one side and at baseband on the other. In addition to the IF and baseband switches themselves, the system comprises monitors of various sorts, logic circuitry at each end of the switching section, communication facilities between ends over the auxiliary radio channel, and manual controls. In contrast to the rest of the TH system, which uses mostly electron tubes, the switching system employs solid-state devices entirely.

I. GENERAL CHARACTERISTICS

The protection switching system permits the replacement of a regular channel by an equivalent stand-by or protection channel. This is done automatically for fades exceeding 27 db in any repeater section and for equipment failures. It can be done manually to release the regular channel for maintenance.

The reliability objectives for the over-all telephone system are that the outage time of a channel should not exceed 0.01 per cent per year, or 0.05 per cent in the worst fading month in a 4000-mile circuit. Several factors like frequency spacing between microwave channels, the number of channels in TH, maintenance procedures, and the length of the switching sections make it necessary to provide two protection channels in TH to meet these objectives.

Protection against fading is possible because fading in microwave systems is generally frequency selective and tends to affect only one broadband channel at a time. A protection channel is normally available when a regular channel fades, and the protection switching system then acts as a switch-type frequency diversity system. However, with only 30-mc separation between radio channels, the probability of two adjacent channels fading simultaneously cannot be neglected

To perform maintenance on equipment, the channel involved is replaced by a protection channel, which makes the latter unavailable for other channels which might get into trouble. The need for two protection channels is again clearly indicated. Daytime maintenance of TH is recommended to avoid the period of heavy fading during the night and especially during the early morning hours. The longer the switching section, the higher is the probability of fading or equipment failure on a given channel, and the more difficult it becomes to meet the reliability objectives. Accordingly, switching sections contain at most ten repeater sections; the average is six.

Table I gives calculated channel outage times for a 4000-mile TH system, taking into account fading, equipment failures, daytime maintenance, and an average switching section length of six repeaters. The table shows that a single regular channel protected by only one protection channel does not meet the outage time objectives given above. The calculated outages, however, are not serious enough to warrant the

Table I — Calculated Outage Times for TH Automatic Protection Switching*

For 4000-mile route due to fading, daytime maintenance, and equipment failures. Average switching section length of six repeaters.

Channels Equipped		Worst Month %	Annual %
Regular	Protection		
1	1	0.111	0.038
2	1	0.167	0.057
3	1	0.222	0.076
2	2	0.0016	0.0005
3	2	0.0040	0.0011
4	2	0.0078	0.0019
5	2	0.0130	0.0033
6	2	0.0198	0.0049
Objectives		0.05	0.01

* Table by Mr. J. F. Laidig.

installation of two protection channels with the first regular channel. As soon as the second regular channel is installed, two protection channels are provided, and the outage time objectives are met even with the ultimate complement of six regular channels. It is of interest to note that in a system without protection switching, a broadband channel would be out of service several hundred hours per year due to fading and failures alone, not considering maintenance.

In addition to the outage time objectives, there are systems requirements imposed by the need to avoid transmission disturbances, called hits, produced by fading or maintenance switching. The requirements on hits vary with the type of service carried over the broadband channel — telephone circuits being more tolerant than telegraph or data. Switching hits can be caused by shorts, opens, misterminations or transient voltages produced by the switches themselves during the switching interval. The requirement on this is about 10 microseconds for the transient duration. Field tests show that the circuit design is such that these hits do not affect any services presently transmitted over TH.

A second type of hit is due to level or phase differences between the two channels involved in a switch. The objectives for this are that the instantaneous level change should not exceed 0.25 db at any frequency in baseband, and that the signal shift should not exceed 10 millimicroseconds. This latter requirement means that the absolute transmission time over a switching section (which could be over 300 miles) has to be the same for all eight channels within about six feet of IF cable. Meeting this objective turns out to be a rather involved feature of TH. Briefly, the differential delay of the channels is measured, and the shorter ones are built out to equal the longest with suitable lengths of IF cable. Every effort is being spent to hold the level and phase differences between channels within acceptable limits.

Another type of interference which is easily overlooked is produced by the removal of circuit units for maintenance. Special attention has been given to this problem, and as a result routine maintenance operations, especially on the switches, do not interfere with the signals carried over the broadband channels.

A transmission interruption cannot be avoided, however, when the signal in a channel suddenly disappears due to equipment failure. It takes the system a certain time to transfer service to a protection channel. The signal outage is from 5 to 40 milliseconds, depending on the length and the type of switching section. This figure is well below the 50-millisecond maximum set by seizure of the common equipment in a telephone dial office.

II. THE TH SWITCHING SECTION

As shown in Fig. 1, the switching sections on a TH route form an unbroken chain. Several types of switching sections are possible. The most common are IF-to-IF over several repeater stations and intra-office (local) baseband-to-IF (and IF-to-baseband) around FM terminals. Other types of switching sections are also possible, like the baseband-to-IF sections extending over radio repeaters shown in Fig. 1 and the very rare case of baseband-to-baseband switching.

Each switching section is completely self-contained and has no interconnection with other sections. A switching section protects both directions of transmission. The switching equipments for the two directions are essentially alike but independent of each other. Fig. 2 shows the block diagram of a generalized TH switching section for one direction of transmission. The different blocks of Fig. 2 are described in the following paragraphs.

2.1 *TH Broadband Transmission Facility*

This block represents different possible broadband transmission arrangements as, for instance, a microwave link with several repeater stations starting with radio transmitters at the input and ending with radio receivers at the output. In this case, ingoing and outcoming broadband signals are at IF (74 mc). For terminal switching only, the transmission facility represents transmitting or receiving FM terminals with inputs and outputs at either baseband or IF.

2.2 *Broadband Channel Monitors*

Associated with the transmission facility of Fig. 2 are monitors which check the broadband channels and produce an output if the channels they are associated with are in trouble. In TH, monitors are located at both ends of a switching section and at intermediate repeaters. In the TD-2 protection switching system[1] only one monitor, sensing the signal-to-noise ratio, is used per channel at the receiving end of a switching section. This method cannot be used in TH because a very noisy signal is replaced by an automatically inserted IF carrier.[2]

There are several different IF monitors, which sense the amplitude of the IF carrier. The monitor in the radio receivers operates from the automatic gain control (AGC) system of the main IF amplifier[2] and produces an output when the IF carrier drops approximately 27 db below normal, for instance, due to fading in the microwave path. For a sudden loss of carrier, this AGC monitor reacts relatively slowly (within about

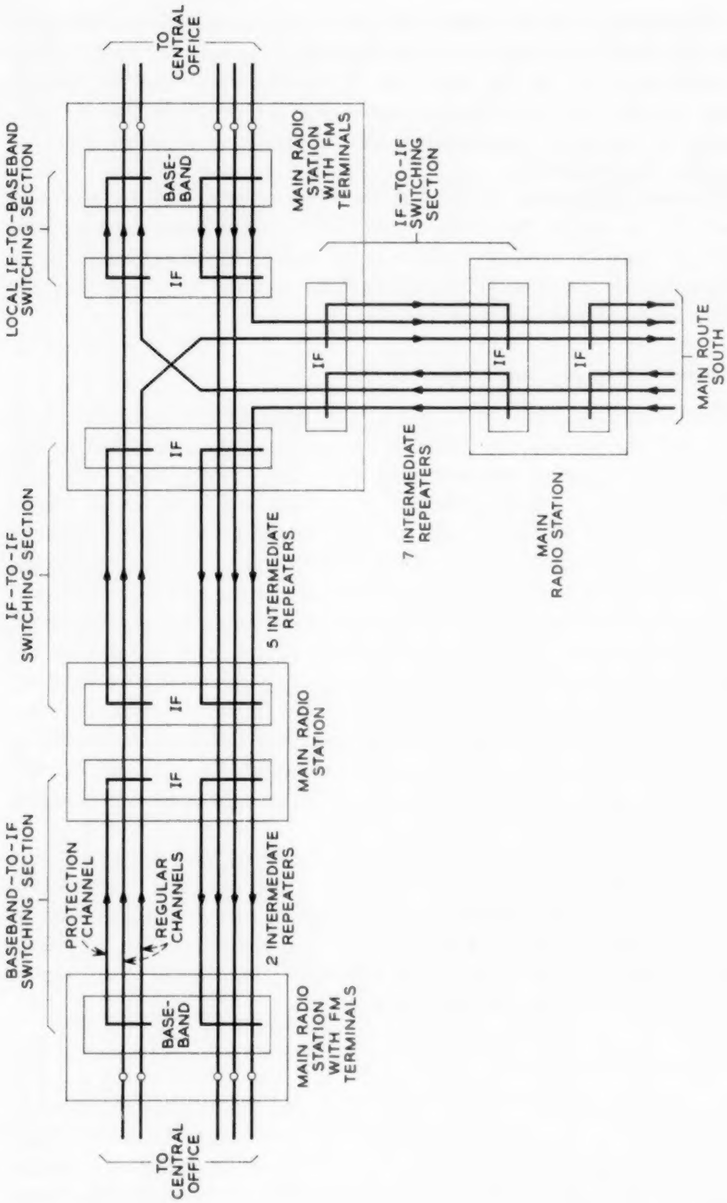

Fig. 1 — Types of TH switching sections.

7 milliseconds). A 2-millisecond response to equipment failure is obtained from the carrier resupply circuit of the radio transmitter.[2] Still another IF monitor is used at the very end of a switching section to indicate failure of the active equipment preceding the switches. In the FM receiver, a voltage proportional to the IF carrier is obtained from the frequency discriminator.[3]

Baseband equipment in the FM terminals is protected by monitors which operate when the dc operating points in the balanced baseband amplifiers have changed beyond a certain limit.[3] This monitoring scheme is considerably simpler than a pilot-tone method but only slightly inferior in its ability to detect trouble.

2.3 *Transmitting and Receiving Switches*

When a switch is requested, the regular channel is first bridged to a protection channel at the transmitting end and then switched to the protection channel at the receiving end. The transmitting switch is bridged across the incoming regular channels A to F as shown in Fig. 2. The signal on any of the six regular channels can be connected to either of the two protection channels, designated X and Y, without producing hits on the through-going channel. The transmitting switches at IF and at baseband are diode switches which operate within a few microseconds.

In the case of a fading or maintenance switch, the transfer at the receiving end occurs between two channels carrying essentially identical signals. A hitless transfer is ensured at IF by the use of diode switches which operate in 2 microseconds. The receiving baseband switches use relays. No transmission disturbances are caused, however, during the one-millisecond transfer interval due to the use of a hitless bridging method, as described in Section 4.3 below.

Not shown in Fig. 2 are minor subsidiary circuits. Associated with the transmitting IF switch are two oscillators which inject IF carriers into the protection channels when they are in the stand-by condition. Special access switches are provided in the protection channels at IF and baseband, to permit use of these channels as additional regular transmission channels under emergency conditions.

2.4 *Signaling Facility*

The special signaling facility (Fig. 2) is used to transmit the monitor indications to the transmitting end and the receiving switch orders from the transmitting to the receiving end of the switching section. The major portion of the logic for the switching system is located at the trans-

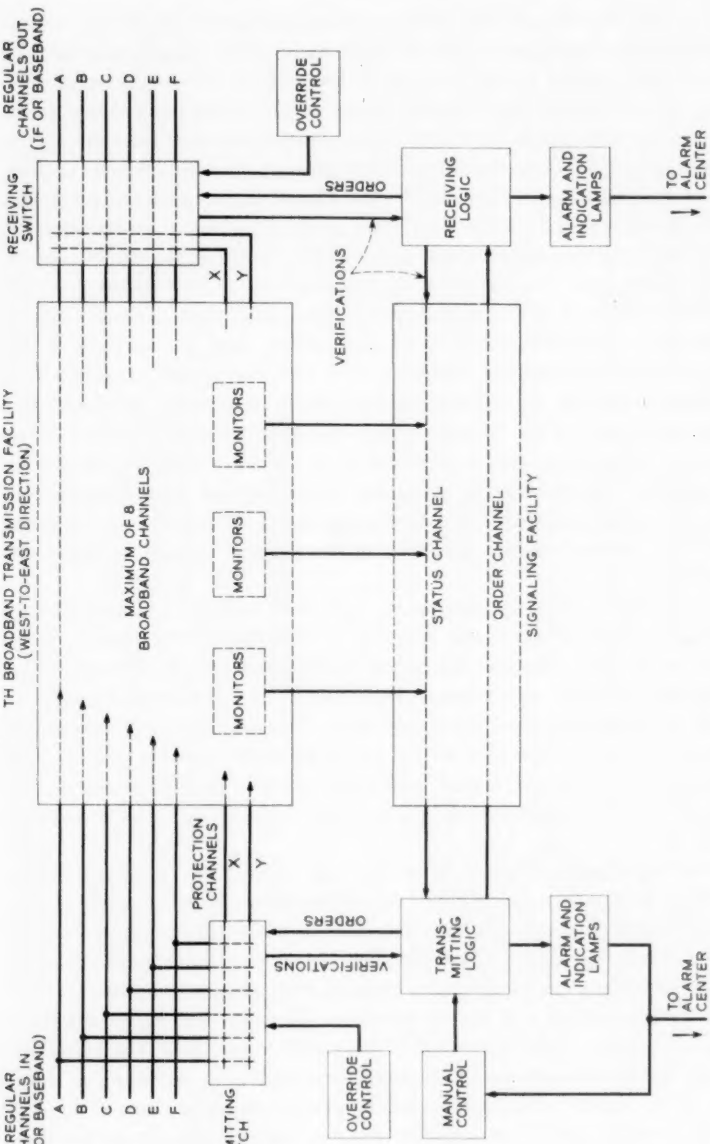

Fig. 2 — Block diagram of TH switching section.

mitting end. Whenever a switching section includes at least one microwave hop, the auxiliary radio channel[4] handles the signaling, which is then done by means of tones in the 20-kc to 36-kc frequency band.

Fig. 3(a) is a detailed block diagram of the signaling facility using tones as the information carriers. The monitors connect into the status channel, which has a maximum of eight tones, one per broadband channel, at 2-kc spacing from 21.5 kc to 35.5 kc. These status tones are produced at the receiving end of the switching section by the tone-transmitter circuit and are detected at the transmitting end by the tone-receiver circuit. Note that the direction of transmission of the status tones is opposite to that of the broadband channels which they are monitoring. Under normal conditions the tones are present, and the outputs of the tone receiver are positive voltages. If a monitor reports trouble in a broadband channel, the corresponding tone is suppressed as long as the trouble persists and the voltage in the output of the tone receiver drops to zero. Status tones can be suppressed by the tone-reporter circuits in the repeater stations or by operating a gate in the tone transmitter. Monitors at the transmitting end of a switching section do not suppress tones but instead operate directly on the output voltage of the tone receivers.

Two groups of four tones each are used for ordering the operation of receiving end switches. These order tones are spaced 2 kc apart, from 20.5 kc to 34.5 kc. The first four tones control the transfer of any of the six regular channels to protection channel X, and the second group of four tones is used for protection channel Y. The ordering code consists of interrupting two of the four tones, which gives six possible orders. The redundancy of the two-out-of-four code allows error checking to be employed on the receiving side with a considerable degree of immunity from noise.

In the signaling facility the West-East and East-West directions of the switching section are intermixed. A maximum of 16 interleaved status and order tones belonging to two different directions of broadband transmission travel together in the same direction over the auxiliary channel.

If the switching section includes only an FM terminal, the two ends of the switching section will not be separated by more than a few hundred feet and normally will be located in the same room. Now, the auxiliary channel is unnecessary and the protection switching signaling is done by on-off dc signals which are transmitted over separate wires (Fig. 3b). The dc reporter circuit accepts the monitor outputs and drops the normally positive output voltage to zero whenever a monitor indicates trouble. With wire interconnections, coding is not needed in the order channel.

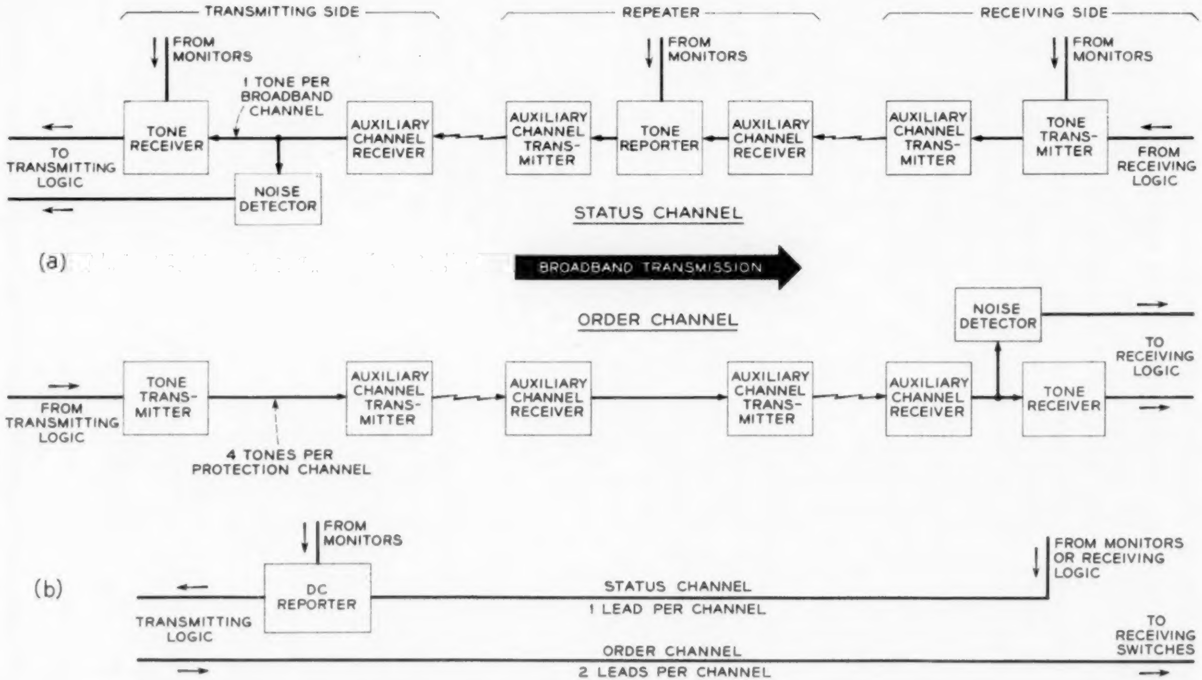

Fig. 3 — Block diagrams of (a) tone signaling and (b) dc signaling.

2.5 *Transmitting and Receiving Logic Circuits*

The main logic circuit is at the transmitting end of the switching section. With monitors in all repeater stations, on the average a switch request originates at the middle of the switching section. Because the transmitting switch always has to be operated first, the speed of operation is increased by sending the status information directly to the transmitting end of the switching section. The transmitting logic receives its orders from the monitors over the status channel or directly from a manual control circuit. The action of the logic circuit depends on the state of the switching section at the moment a switch request is received. If a protection channel is available, the logic orders the appropriate transmitting switch directly and the receiving switch over the order channel of the signaling facility. The transmitting logic also releases the switches after a channel has become good again. After ordering or releasing transmitting or receiving switches, the logic always waits for verification signals regarding the operation of the switches. The logic handles one request at a time. Due to the high speed of operation, such sequential operation is completely adequate.

If a broadband channel fades or fails and no protection channel is available, the logic issues an alarm indicating a loss of service. The transmitting logic also recognizes other trouble situations which may arise within the switching section. If a switch lasts longer than 45 seconds, it is most likely caused by an equipment failure and not fading, and an alarm is therefore produced. Failure to release a switch because of faulty switching equipment and the loss of all status tones due to an auxiliary channel failure also produce alarms. In case of status-channel failure, the logic maintains all previous assignments. If the carrier of a broadband channel disappears at the input of the switching section, switching is inhibited by the transmitting logic because the trouble condition exists ahead of the switching section.

The two-out-of-four code used for ordering the receiving switches in the case of tone signaling is generated in the transmitting logic and is decoded at the other end in the receiving logic. The receiving logic decodes and error-checks the incoming order and in turn sends out signals to operate the receiving switches. The receiving logic maintains the status quo of the receiving switches and issues an alarm if an ordering code containing a single error is received.

2.6 *Manual Control*

The manual control circuit works directly into the transmitting logic (Fig. 2). Two operations are possible. The first is a normal switch

operation which transfers the signal from a regular channel to a protection channel. The manual switch control simulates a channel failure, and the logic processes this request as if it were a legitimate one. Such a manual switch makes a regular channel available for maintenance. The second manual operation is channel lock-out. The regular channels are locked out by making them appear good to the transmitting logic, regardless of the actual situation. Conversely, a protection channel is marked bad for lock-out. The locked-out regular or protection channels, therefore, are essentially removed from the switching system, and the regular channels in particular are unprotected. Lock-out is mainly used during maintenance operations. If a regular channel is temporarily carrying no signal, its lock-out results in better protection for the other channels. A protection channel is locked out if it is used as a regular channel for emergency make-good of a failed transmission facility not normally associated with the TH system. For this purpose, the protection channels are accessible through special access switches.

A set of lamps at the transmitting and receiving ends indicates the status of the switching system. Lamps indicating channel failure, switch assignments, manual switches and alarms are provided.

Manual control is also possible from a remote location. The C1 alarm system is used for the transmission of manual orders from the remote location to the switching station and of status information in the reverse direction for the operation of indicator lamps at the remote location.

2.7 *Override Control*

The override control circuits shown in Fig. 2 permit direct operation of the transmitting and receiving switches. The connection between the switches and the logic is thereby interrupted, and all automatic protection is lost. Previous assignments, however, are maintained by the override control. The override switches are used mainly when serious trouble in the logic circuits calls for taking them out of service for trouble-shooting and repair. This rather radical method is necessary because large parts of the logic are common to all channels and cannot be maintained on a per-channel basis.

No remote control is provided for the override switch control; thus service personnel must coordinate operations at both ends of the switching section. An override operation at one end of the switching section always maintains status quo at both ends. This is accomplished automatically in the case of tone signaling by interrupting all tones going to the other end. The loss of all tones makes the logic circuits inoperative, but previous assignments are maintained. For dc signaling only the transmitting override control circuit is needed, since this can operate

transmitting and receiving switches simultaneously over wire connections.

III. THE AUTOMATIC SWITCHING OPERATION

The switching system, headed by the transmitting logic, can perform four main, different automatic switching operations. In the following sections the sequence of events is described, starting with the appearance of the initiating signal and ending with the completion of the automatic process.

3.1 *Signal Transfer from Failed or Faded Regular Channel to Protection Channel*

Trouble in a broadband channel is discovered by a monitor and reported over the signaling system to the transmitting logic. The transmitting logic selects a protection channel and orders the transmitting switch to be operated. If no protection channels are available, a service failure alarm occurs. A receiving switch order is sent out only after the logic has received a switch-verification signal and has checked that the trouble does not originate in the previous switching section. Depending on whether tone or dc signaling is used, the receiving switch is either operated through the decoding receiving logic or directly. The operation of the receiving switch concludes the important part of the switching cycle as far as transmission in the broadband channels is concerned. Depending on the kind of switching section, the time from the appearance of the trouble to the receiving switch transfer can vary from as little as 5 to as much as 40 milliseconds. This is the time the signal would be lost in case of a sudden equipment failure. The signal over a fading channel, however, is switched without interruption.

The operation of the receiving switch causes a steady verification signal to be fed to the receiving logic. This is translated by the logic into a short interruption of the status signal associated with the protection channel involved. This verification signal is then received in the transmitting logic, which recognizes successful completion of the switching operation.

If the verification signals from the transmitting and receiving switches are not received within 65 milliseconds, the transmitting logic tries to switch to the other protection channel. Where only one protection channel is available, the failed regular channel is locked out and a service failure alarm occurs. The lock-out forces transmission back onto the regular channel (see Section 2.6 above). Unlocking then requires a manual operation.

3.2 *Signal Restoral from Protection Channel to Operative Regular Channel*

When the failure or the fade in a channel is over, the monitors restore the status signal of the channel. The transmitting logic first releases the receiving switch over the order channel. After the receiving switch-release verification is received, the transmitting logic resets the transmitting switch, which completes the switch-release operation. The release verification is in the form of a short interruption of the status signal associated with the protection channel involved. If the receiving switch verification does not arrive within 55 milliseconds, the regular and the protection channels involved are locked out by the transmitting logic, and a release-failure alarm is issued. After clearing the trouble, the channels can be unlocked manually.

3.3 *Recognition of Failure in Previous Switching Section*

A loss of IF signal at the input of a switching section is immediately recognized by the first IF monitor in the regular channel involved, and a switch to a protection channel is made at the transmitting end. But the protection channel is also marked bad following the switch operation because there is no signal entering this channel either. The first and the second failure indication follow each other within a few milliseconds, which fact is recognized by the transmitting logic as a preceding section failure. The regular channel then is temporarily (110 milliseconds) locked out by the logic. During this time, other channels have a chance of getting served by the logic. The signal loss at the input of the switching section will in most cases not last longer than 40 milliseconds, namely the time it takes to make good a failure in the last part of the preceding switching section. (Only failures after the last carrier resupply circuit in a switching section cause loss of IF carrier.) After 110 milliseconds, the channel is unlocked. If the signal has not reappeared, the same inhibit check starts over again.

3.4 *Signal Transfer from Failed or Faded Protection Channel to Second Protection Channel*

If a protection channel carrying a switched signal fades or fails, a transfer to the second protection channel is made. This transfer always causes a short signal interruption (10 milliseconds) even in the case of fading, because transmitting-end bridging of two protection channels is not possible. The transmitting and receiving switches of the failed protection channel are immediately released by the logic, and the transmitting bridge to the second protection channel is established.

After waiting for the inhibit check, the receiving switch transfer to the second protection channel is ordered. If no second protection channel is available, no switches will be executed by the system.

The following sections give a more detailed description of some of the important circuits of the switching system.

IV. THE TRANSMITTING AND RECEIVING SWITCHES

The switches must meet the stringent transmission requirements of the broadband channels, provide switching operation which is fast and hitless, and be removable for maintenance without disturbing the signals on the channels.

4.1 *IF Switches and Associated Circuits*

A schematic of the transmitting IF switch is shown in Fig. 4. The switch is bridged across the regular channels by means of IF directional couplers. These couplers have negligible loss in the regular signal path but attenuate the signal to the switch by 20 db. The switches are, therefore, well isolated from the through line without the aid of bridging amplifiers. Directional couplers are needed, rather than resistive pads, to discriminate against the echo returned from the far end of the IF cable leading to the regular channel equipment. The 20-db bridging loss and additional losses through the switch are made up by two IF amplifiers in the switch outputs to the X or Y protection channels.

The gates are represented symbolically in Fig. 4 by boxes containing single-pole, single-throw switches. The signal going over channel A, after entering the switch, passes through the 3×1 interconnecting network and is usually terminated in a 75-ohm resistor through gate A/Z. To switch this signal to protection channel X, switch A/Z is opened simultaneously with switch A/X being closed. Through the 6×1 interconnecting network, the signal passes on to channel X.

The gate circuit, shown in Fig. 5(a), contains three diodes which can be forward or reverse biased by the application of the appropriate switching voltages. Simplified ac equivalent circuits for the ON and OFF conditions are shown in Figs. 5(b) and 5(c). The ON gate represents a matched resistive T pad with a 75-ohm characteristic impedance and 1.5 db of insertion loss. The gold-bonded germanium diodes used in this circuit have a 5-ohm resistance in the forward and a capacitance of 0.8 pf in the reverse biased condition. The loss through the OFF gate is about 90 db at 74 mc. A photo of the unit is shown in Fig. 6.

Since the ON gates have 1.5 db loss, the switch is arranged so that the

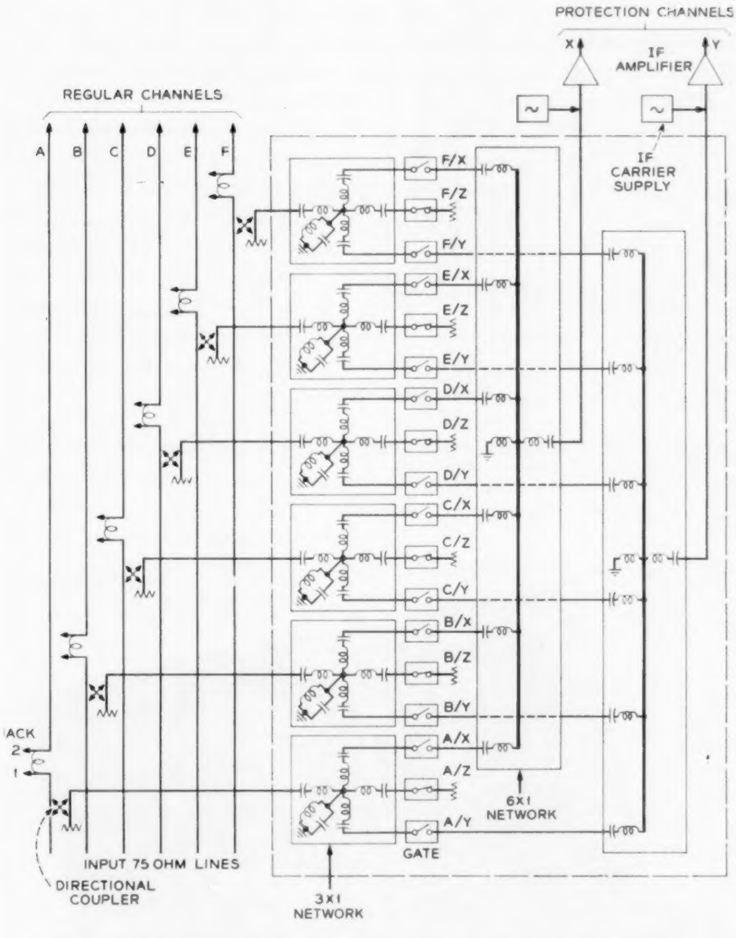

Fig. 4 — Schematic of IF transmitting switch and associated circuits.

IF signal passes through no more than one gate under any switching condition. It is seen from Fig. 4 that the incoming signal on channel A, for instance, after first going through the 3×1 interconnecting network, appears at the input terminals of three gates simultaneously. Only one of the gates is transmitting at a time and furnishing the 75-ohm network termination. The other two gates are OFF and load the network capacitively. The 3×1 network appears as a band-pass network to the

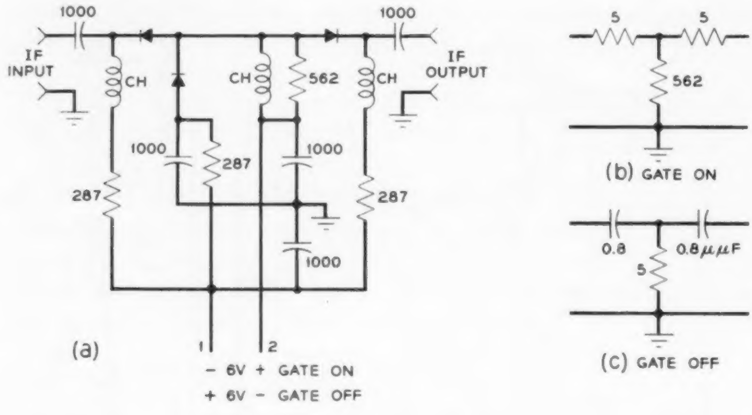

Fig. 5 — The IF gate: (a) circuit diagram, and ac equivalent circuits for the (b) gate ON and (c) gate OFF conditions.

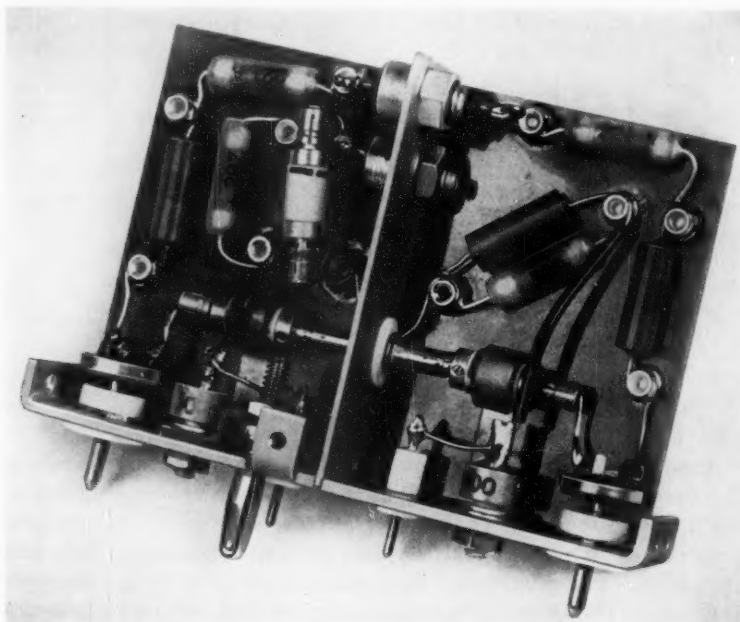

Fig. 6 — Photograph of the IF gate.

transmitted signal. The small capacitance of the OFF gates contributes to the total shunt capacitance of the filter. Over the 58-mc to 90-mc band, the gate capacitance is essentially unaltered by the series LC circuit which connects the gate to the shunt point of the filter. Due to symmetrical physical layout of the circuit, all three paths are electrically identical and have the same transmission characteristic.

Basically the same but somewhat more complicated situation exists at the output side of the gates where connection is made to a 6×1 network for either channel X or Y. There are six possible paths to each protection channel. If channel A, for instance, is switched to channel X, gate A/X is ON, and the other five gates, B/X to F/X, connecting to the 6×1 network for channel X, are all OFF. The same electrical and physical requirements as for the 3×1 network apply also for this 6×1 network. The equivalent circuit for the network and the throughpath is again a band-pass filter.

The IF gates and the 3×1 and 6×1 interconnecting networks are all built and tuned up separately. After they are assembled in an aluminum casting (Fig. 7) the following transmission requirements are met without further adjustments:

Insertion loss at 74 mc	1.9 db
Transmission flatness, 58 to 90 mc	±0.05 db
Return loss, 58 to 90 mc	≧28 db
Isolation between channels at 74 mc	≧86 db

The IF switch of Fig. 4 is equipped with only the minimum number of gates and networks needed for the number of channels installed. It is imperative, however, for a network to be electrically loaded with the full number of gates, or their equivalents, to maintain its transmission characteristic. Network connections to unequipped channels are terminated with the impedance of nontransmitting gates, simulated by passive dummy gates which have the same impedance but require no driving voltages. Fig. 7 shows a partially equipped switch.

An IF gate can be removed for maintenance without interruption or noticeable degradation of the broadband signal. Only a nontransmitting gate can be removed, and only one at a time. By suitable manual switching it is possible to bring any gate into the OFF position. To avoid serious transmission disturbances during removal of a gate, the switching voltage and ground are disconnected only after opening the IF connections to the gate. This is achieved by proper mechanical connector construction. On insertion, the IF connections are made last.

The gates are controlled by switch-control circuits. One control circuit serves the three gates associated with a regular channel, e.g. A/X, A/Y and A/Z. The circuit contains not only transistor amplifiers for the low-level logic signals but also an OR logic circuit for the operation

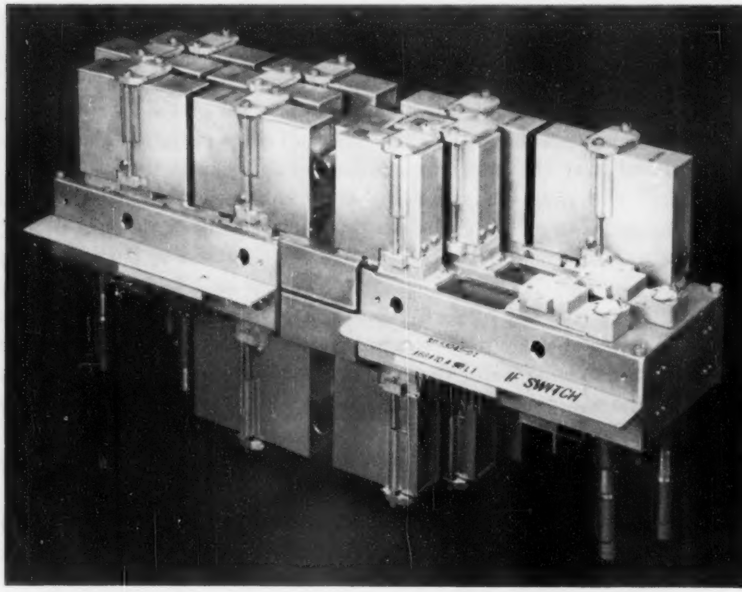

Fig. 7 — Partially equipped transmitting IF switch (4 regular and 2 protection channels).

of gate A/Z. The switching voltages change so fast that the transfer of the IF signal is completed within two microseconds. The presence of the correct operating voltage at the switches is taken as a confirmation that the switches have operated properly.

If no signals are switched over the protection channels, 74.13-mc carriers are injected to prevent the monitors from indicating the protection channels bad. The carriers are produced by the IF carrier supply circuits, which are bridged across the lines to the protection channels as shown in Fig. 4. The oscillators are started and stopped in synchronism with the switch operation through control signals from the transmitting logic. The IF carrier supply circuit is very similar to the IF carrier resupply circuit described in another article[2] of this issue. Its output power, however, is much lower because the carrier is injected at a low level point just ahead of the 22-db IF amplifier.

Specially developed jacks are inserted in the regular channel paths right after the 20-db directional couplers. These jacks give access to the regular channels for mainenance. A spring contact inside the jack provides a continuous path between the two sides of the jack under normal

conditions. The conventional patch plug can therefore be dispensed with. A regular channel is made available for maintenance by first switching the signal over a protection channel and then plugging a 75-ohm termination into side 1 of the jack. The signal flow over the regular channel is now interrupted, and access to this channel is obtained through side 2 of the jack. The make-before-break insertion of the termination on side 1 produces a mistermination on the line during the very short insertion period, but this mistermination is isolated from the signal transmission path by the directional coupler.

The receiving IF switch is similar to the transmitting switch shown inside the dashed area of Fig. 4, except that the signals pass through the switch in the opposite direction. The Z gates, A/Z etc., are now connected to the outputs of the radio receivers instead of to terminations. The protection channels are not terminated by the switch under normal conditions, and the regular channels lose their termination when switched. This mismatch does not affect the transmission over the system, and the level-sensing monitors, which are located ahead of the receiving switch to supervise active equipment ahead of the switch, are isolated from the resultant reflected waves by directional couplers.

Like the transmitting switch, receiving switch transfers are completed within two microseconds. If the absolute delays of the regular and protection channels are equal, none of the transmitted services is affected by this two-microsecond transfer.

4.2 *Transmitting Baseband Switch and Associated Circuits*

The transmitting baseband switch with its associated circuits is shown in Fig. 8. The incoming baseband signals (e.g., from the multiplex telephone terminals) are normally connected through 0-db bridging amplifiers A to the FM transmitters. They can also be connected to the protection channels through the switch when necessary. The switch is composed of a number of diode gates and is quite similar to the IF switch with the exception that it is arranged in two stages. Any of the regular channels can be transferred to protection channel X through the first stage of the switch and to protection channel Y through the first and second stages in tandem. Baseband switches are generally used in the very short switching sections extending around a terminal, where one protection terminal is usually sufficient and only the first stage of the switch is needed. The two-stage switch therefore has the advantage of greater circuit economy over the single-stage switch used at IF.

The amplifiers A are bridged across the 124-ohm balanced line with a high impedance, and the termination for the line is provided through

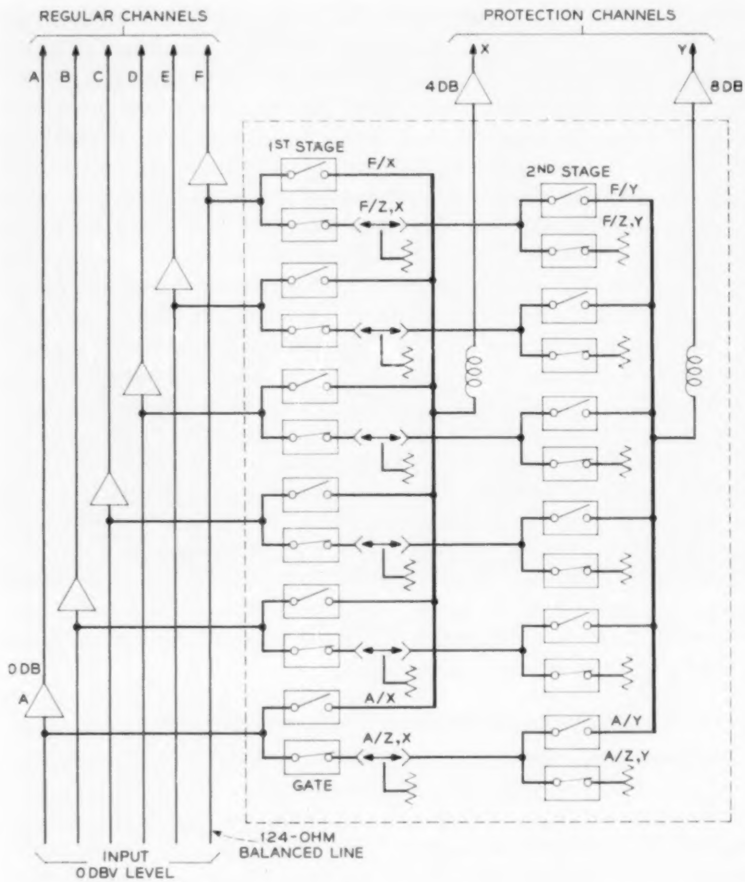

Fig. 8 — Schematic of transmitting baseband switch and associated circuits.

gates A/Z, X and A/Z, Y. If channel A is switched to channel X, for instance, the A/X gate is made transmitting simultaneously with the A/Z,X gate being made nontransmitting. The switching process requires 10 microseconds, and during this time the line is incompletely terminated. However, none of the services transmitted over TH is impaired by this effect. The networks used to couple the gates to the incoming line or the protection channels are low-pass filters, which are much simpler than the interconnecting networks in the IF switch.

The baseband gate must meet close transmission, return-loss, and

isolation requirements, as well as very stringent requirements on the voltages which may appear on the line during and after the switching operation. The balanced 124-ohm baseband gate uses twelve diodes in a doubly symmetrical configuration as shown in Fig. 9(a). Figs. 9(b) and 9(c) show the dc equivalent circuits for the gate in the ON and the OFF state. If the driving voltages for the gate are exactly symmetrical to ground (±10 volts), less than 5 millivolts will appear at terminals 1, 2, 3 and 4 to ground. No adjustment of components is required to obtain this low pedestal voltage.

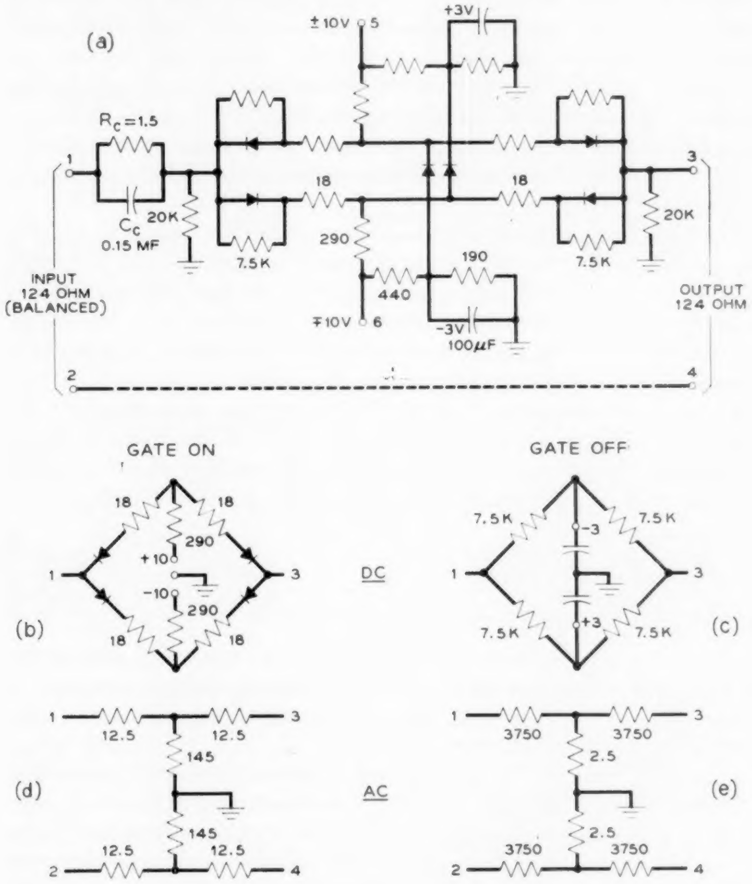

Fig. 9 — The baseband gate: (a) circuit diagram and equivalent dc and ac circuits (b through e).

The gate control voltage is produced in a special switch control circuit which receives its orders from the transmitting logic. One such circuit is needed to operate the pair of gates associated with a particular regular and protection channel, e.g., the pair A/X and A/Z,X. Special circuitry holds the unbalance of the ±10-volt gate control voltages within 0.2 volt, even during the 10-microsecond transition to ±10 volts.

Taking the slight assymmetry of the switching voltage into account, there never appears more than 25 millivolts from any gate terminal (1, 2, 3 or 4 of Fig. 9) to ground. A large part of this voltage appears as a longitudinal component on the 124-ohm balanced line and therefore does not interfere directly with the transmitted signal. Television, which passes through the switches at a −12-dbv level, is the service most susceptible to switching pedestals and transients. The TV interference becomes invisible if the symmetrical voltage is below 15 millivolts. Tests show that most gates in a baseband switch cause no disturbances at all to a TV picture, with some producing a barely visible streak across the screen.

The gates are designed to transmit frequencies from dc to 10 mc. The ac equivalent circuits are shown in Figs. 9(d) and 9(e). In the transmitting or ON state, the gate acts essentially as a matched resistive pad with approximately 4 db loss. This loss is compensated by a 4-db amplifier in channel X and by an 8-db amplifier in channel Y. Transmission flatness is improved by the addition of resistor R_c and capacitor C_c. These components equalize a 0.05-db transmission drop around 1 mc, which is due to the inductive character of the forward-biased diodes. The diodes are the same gold-bonded germanium diodes used in the IF switch. The transmission characteristics of the gates are as follows:

Insertion loss at 5 mc	4 db
Transmission flatness, dc to 10 mc	±0.02 db
Return loss, dc to 10 mc	≧38 db
Modulation products	−70 db
Isolation between input and output, dc to 10 mc in OFF state	≧86 db

As in the case of the IF switch, dummy OFF gates are used in the positions of the 6 × 1 interconnecting network which correspond to channels not installed. The general equipment arrangement is also similar to the IF switch.

The 0-db, 4-db and 8-db baseband amplifiers all use the same basic circuit (Fig. 10). They are balanced, transistorized amplifiers and differ only in the amount of feedback. Each side of the amplifier consists of three transistor stages using germanium diffused-base transistors with an alpha cutoff frequency of about 750 mc. Feedback is provided through resistor R_f which stabilizes the voltage gain of the amplifier to approxi-

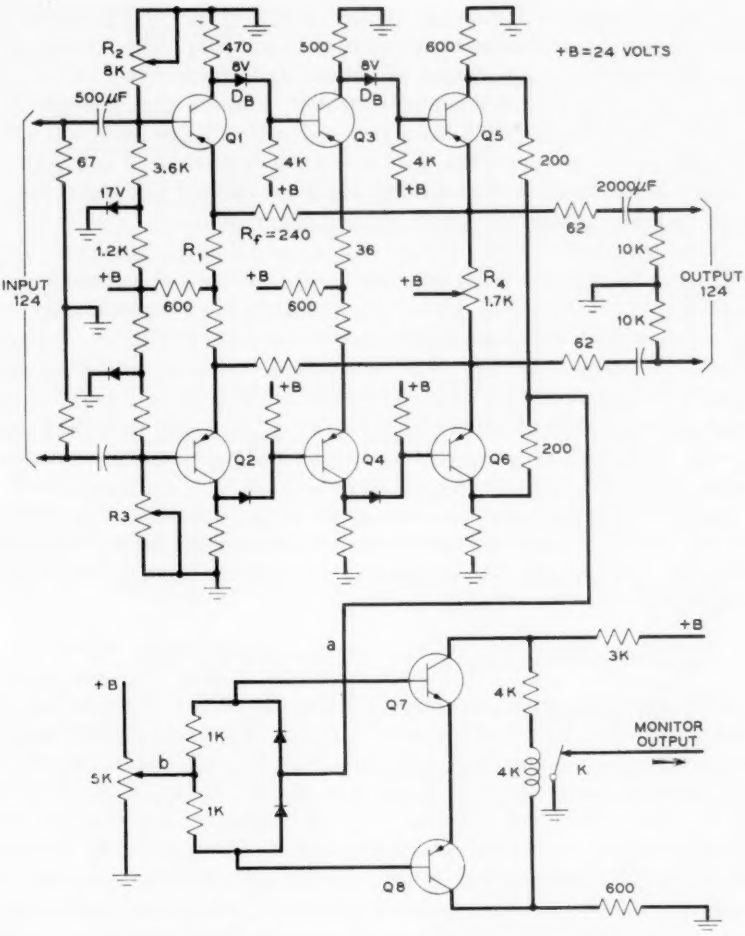

Fig. 10 — Simplified schematic of baseband amplifier.

mately $g = 0.5\ (1 + R_f/R_1)$. The different gains are obtained by proper selection of resistor R_1. The dc coupling of the amplifier by means of voltage breakdown diodes D_B avoids the use of large coupling capacitors and ensures stability of the feedback loop at low frequencies. The amplifier is coupled to the balanced 124-ohm lines on the input and output sides through large electrolytic capacitors which provide flat transmission down to a few cycles. Second-harmonic distortion in the ampli-

fier is minimized by balancing voltages and currents in the amplifier through adjustment of potentiometers R_2, R_3 and R_4. The second and the third harmonics are at least 60 db and 80 db respectively below the fundamental for an output level of 0 dbv. Transmission through the amplifier is flat to ±0.03 db from 50 cps up to 10 mc. The input and output return losses against 124 ohms are better than 40 db. Longitudinal voltage components of the incoming signal encounter much higher feedback in the amplifier and are attenuated about 45 db.

Fig. 10 also shows the circuit used to monitor the amplifier. The voltage at point "a" is not affected by the signals going through the amplifier but is affected by almost any change in the dc operating points in the amplifier. The voltage "a" is compared with a reference voltage "b" in the bridge consisting of two diodes and two resistors. The bridge output voltage is applied to a differential amplifier using transistors Q7 and Q8. If the difference between voltages "a" and "b" exceeds about ½ volt, relay K, which normally is operated, changes state and marks the channel in which the amplifier is connected bad.

Special attention has again been given to the maintenance problem. Gates, baseband amplifiers and switch control circuits can be removed from and placed into the equipment bays without causing hits in the channels.

4.3 *Receiving Baseband Switch*

The receiving switch uses relays for the sake of low insertion loss, a choice which is feasible because of the possibility of a hitless signal transfer at the receiving end. A complete switch for six regular channels and two protection channels is shown in Fig. 11. It consists of 24 wire spring relays which can easily be cascaded because the loss through any of them is only a few hundredths of a decibel. To simplify Fig. 11 (and later Fig. 12), only one side of the balanced 124-ohm line is shown. A transfer from channel A to protection channel X, for instance, is made by first operating relay R1. The signal on channel X then appears at relay R2, which is operated next. In case of a maintenance or fading switch, the signals on X and A are identical and the transfer by relay R2 can be made hitless as follows. Phase 1 of Fig. 12 shows the relay R2 in its normal or de-energized position. The relay is then operated, and after approximately 10 milliseconds contacts 3 and 4 close (Phase 2). Under the assumption that the TH channels are equalized for absolute delay and equal level, the signals from X and A are identical, and the bridge provided by R2 will not change the output voltage. Note that as soon as either contact 3 or 4 is made, all four arms are connected; hence it is

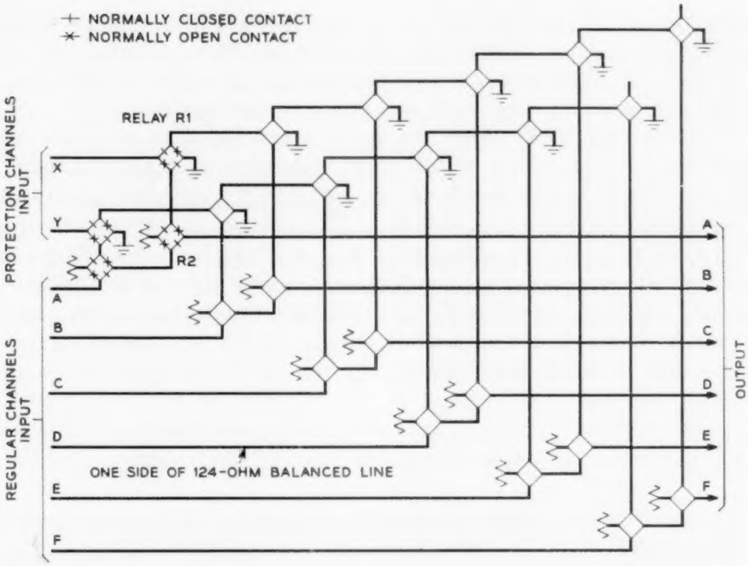

Fig. 11 — Receiving baseband switch for 6 regular and 2 protection channels.

immaterial which makes first or whether they both make contact simultaneously. After a period of approximately one millisecond, contacts 1 and 2 open and the hitless transfer from channel A to X is completed (Phase 3).

The two-step switching operation is necessary to provide enough isolation between regular and protection channels in the normal or un-

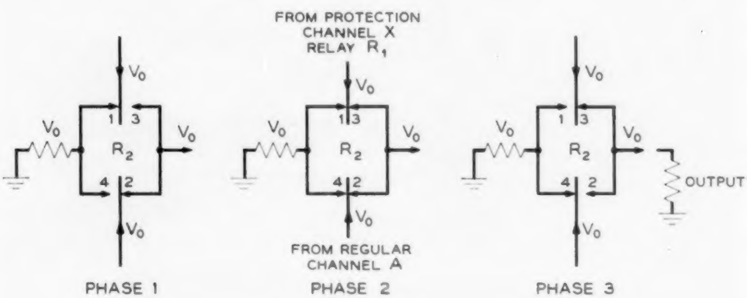

Fig. 12 — The three phases of a hitless signal transfer in the receiving baseband switch. Only one side of the balanced 124-ohm line is shown.

operated condition. The open relay contacts of R1 and R2 individually provide only 45 db of isolation. Cascading two open contacts in R1 and R2 with the connection between the contacts grounded provides more than 90 db of isolation at 10 mc, giving a comfortable margin over a requirement of 86 db. When channel A is switched to X, however, only the parallel capacitance of the single open contacts 1 and 2 (Fig. 12, Phase 3), and therefore only 45 db, separates the switched signal from the voltages appearing on channel A. The isolation is increased from 45 db to 60 db by the use of two pairs of neutralizing capacitors. These are connected from contacts 1 and 2 on each side of the 124-ohm balanced line to contacts 3 and 4 on the other side and thus introduce coupling of opposite phase. Isolation of 60 db is tolerable for this type of exposure because it is much less frequent than the one between protection and regular channel.

The two relays R1 and R2 are contained in a switching unit (Fig. 13) together with a transistorized control circuit. The transmission path in

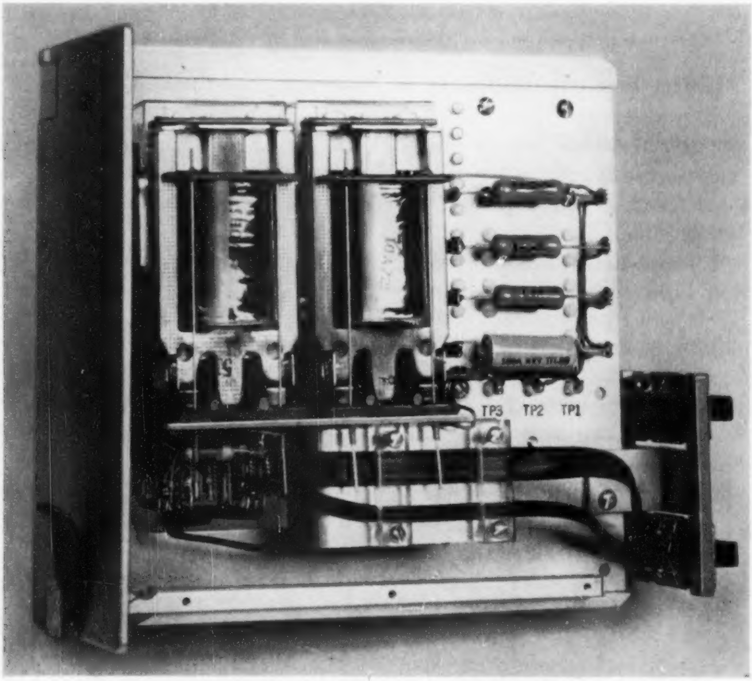

Fig. 13 — Photograph of receiving baseband switch.

the unit extends from a conventional multipin connector through the relays and back to the connector. The excess capacitance of the relays and connectors is compensated and the return-loss requirement of 40 db is met by making the connection with suitable lengths of 150- and 300-ohm "twin lead" cable, which acts as a small series inductance.

The receiving baseband switch, like the other two types, grows with the number of broadband channels installed. The addition of channels can be made without interference to the channels already installed. A switch unit can also be removed from the bay without disturbing service. After manual lock-out of the regular and protection channels involved, the initial small movement of withdrawal of the unit establishes a path in parallel with contacts 2 of relay R2.

V. THE SIGNALING FACILITY

The tones which carry status and order information over the auxiliary channel are produced in the tone transmitter circuit (see Fig. 3a) which consists of up to 16 separate oscillator circuits. At each end of a switching section, a maximum of eight oscillators is needed for the status channel of one direction of broadband transmission and a maximum of eight for the order channel of the other direction of broadband transmission. Due to the high output impedance of the tone oscillator circuits, they can be connected together in any number up to 16 across a 135-ohm unbalanced line leading to the auxiliary channel transmitter. The heart of the circuit is a Colpitts CW oscillator using one transistor. Each oscillator in the 20.5-kc to 35.5-kc range has a long-term frequency stability of ± 10 cycles over wide ranges in temperature and supply voltage. The tone is applied to the output through a gating circuit which, when operated, drops the normal tone level of -18.5 dbm on the 135-ohm line approximately 25 db. To avoid disturbing switching transients in the output, a balanced gate is used. This gate is operated by different input signals. The status-tone oscillators, for instance, can be keyed individually by the broadband monitors located in the receiving station or as a whole by operation of the receiving override switch. The status-tone oscillator associated with a protection channel can be interrupted for about 5 milliseconds by a receiving switch-verification signal coming from the receiving logic.

To reduce the interference into adjacent tone channels, which are only 1 kc away, the spectrum of the keyed tones is restricted by lengthening the rise and fall times to 1 millisecond. This is accomplished by circuitry in the balanced gate and the driver stage. The oscillator is supervised by a comparator circuit which checks whether the incoming voltage and the outgoing tone are compatible.

The tone-reporter circuits are used in the repeater stations to interrupt the status tones associated with channels faded or in trouble. The circuit consists of eight series resonant circuits tuned to the status-tone frequences. When an LC circuit is connected through a relay contact across the 135-ohm line in the auxiliary channel, the corresponding tone is suppressed 18 db; adjacent tones are only slightly affected.

The tone receiver circuit consists of a maximum of 16 detector circuits matching the tone oscillators at the other end of the switching section. Any number of tone detectors up to 16 can be connected across the 135-ohm line coming from the auxiliary channel receiver. The input circuit of each detector consists of a series resonant circuit which is connected to ground through the low emitter-base impedance of a buffer amplifier transistor. The filter represents a 135-ohm load for the tone it is tuned to and a very high impedance for all the other tones. The interaction between circuits therefore is very small. Additional filtering is provided in the collector circuit of the buffer amplifier transistor. The over-all 3-db bandwidth is about 400 cycles, which bandwidth leads to approximately two milliseconds delay of the keyed tone. After filtering, detection and amplification, a Schmitt trigger circuit provides positive indication that the tone has dropped below or risen above a specified level. The tone-detector output drops to zero when the tone level falls 12 db below normal. There is a 2-db protective hysteresis between the switch-off and the switch-on levels so that the tone has to rise to within 10db of normal to restore to a regular +8-volt output signal. These switching levels were chosen to take care of possible gain variations in the auxiliary channel under extreme conditions. In the tone-off state the low switching level makes the detector susceptible to accidental triggering by the fairly high auxiliary channel noise level. In the case of the status tones, such noise makes a failed channel look good temporarily. The protection switch is then dropped with the result that the channel is lost for at least the duration of the noise burst. To counteract this troublesome situation, an unsymmetrical delay circuit is built into the tone-detector circuit. Negligible delay is introduced when the tone disappears. The tone, however, has to be present at least 15 milliseconds before the tone-detector output rises to +8 volts. This delay is not detrimental to the normal operation of the system. Noise peaks lasting this long seldom occur. Measurements show that the number of accidental switches due to noise decreases about six orders of magnitude when the delay is increased from zero to 15 milliseconds.

If the noise from the auxiliary channel should become very high, false switches and service outages would happen quite frequently. It is then

better to disconnect the switching system from the auxiliary channel. This is done by the noise-detector circuit, which connects directly into the transmitting or receiving logic where the status quo of the switching system is maintained. No further switch requests can be served in this case, however. Taking all the conflicting effects into account, use of the noise-detector leads to an over-all reduction of channel outages.

VI. THE LOGIC CIRCUITS

The purpose of the logic and the part it plays in controlling the various switching operations were given in general terms in Sections 2.5 and III. This section describes the constituent parts of the logic and their organization and also presents the major logic diagrams and basic circuits employed.

6.1 *Organization of Transmitting Logic*

A block diagram of the transmitting logic is shown in Fig. 14. The basic action circuits are represented by blocks in reverse type, while those performing control and checking functions are in regular type. The description specifically illustrates the use of the logic in an IF switching section in which tone signaling is employed between the transmitting and receiving ends. To facilitate the study of signal flow, the channel-status leads (Section 2.4) are shown coming into the transmitting logic on the left side, while all outgoing leads are shown on the right side. The status leads associated with regular broadband channels A through F are designated BS1 through BS6, while those associated with protection channels X and Y are designated BS7 and BS8.

The channel-status and memory circuits receive information not only from the status channel of the signaling facility but also from the manual switch control circuit and from points within the logic itself. The various situations encountered by the logic and, especially, the problem of keeping switching operations at both ends of the section properly synchronized make it necessary that the information outgoing from a status circuit, at times, be at variance with the true status of the radio channel it represents. In other words, it is sometimes necessary to make the status of the channel look "good" when it is, in fact, "bad" and vice-versa. The channel-status circuits, accordingly, include a flip-flop memory in addition to a system of gates in order to make possible the required controls.

When a regular transmission channel fades or goes "bad", the channel-status circuit presents a signal to the assignment circuit calling for the assignment of one of the protection channels as a substitute, provided,

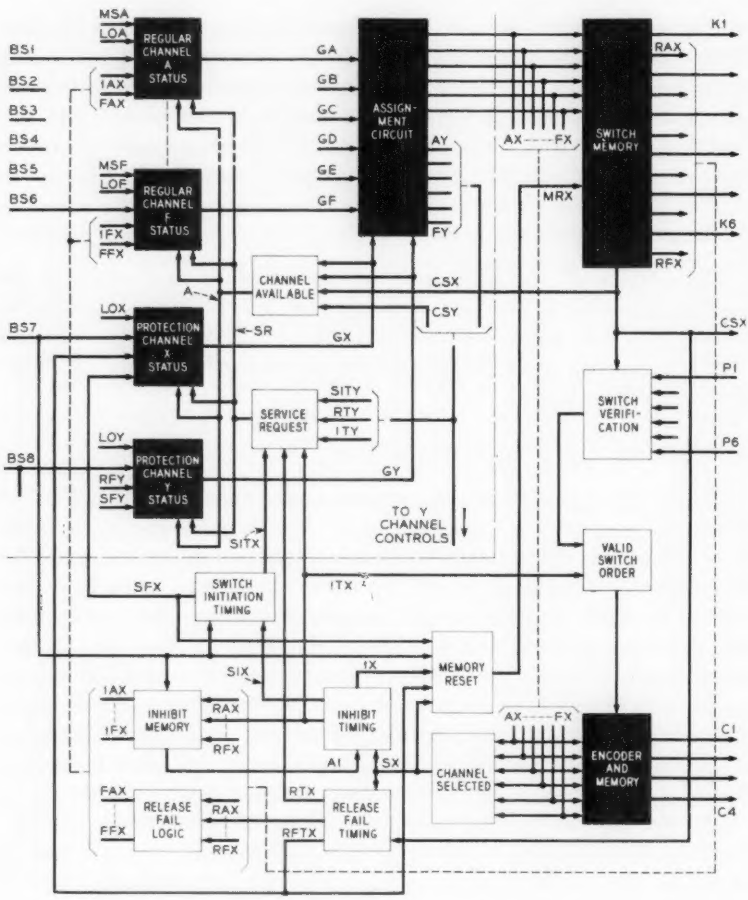

Fig. 14 — Block diagram of transmitting logic.

(a) there is at least one protection channel available for use and (b) the logic control and checking circuits are not busy serving some previous request for service. For a protection channel to be available, its transmission must be good, as indicated by the voltage level on the GX or GY lead, and it must not already be in use, as indicated by the voltage on the CSX or CSY lead. If at least one protection channel satisfies these conditions, the channel-available circuit makes this fact known to the channel-status circuit via the A lead. The service-request circuit, via the SR lead, can inhibit the generation of any new output signals in the

channel circuits during the interval in which a previous request is being served. A request for service on the part of a channel, when permitted by A and SR lead indications, is made known to the assignment circuit by a change in voltage on one of the leads designated GA to GF. A signal on the GA lead, for example, if protection channel X is available, results in a signal outgoing from the assignment circuit on the AX lead. This leads into the next stage of the work cycle.

The circuits involved in the assignment function are common to both protection channels. The remainder of the circuits, however, are individual to a given protection channel. Thus, the six assignment leads designated AX to FX go to circuits associated with channel X, while the six leads designated AY to FY go to circuits associated with channel Y. In Fig. 14 the blocks below and to the right of the dashed line represent only channel X control circuits.

Immediately after the assignment signal is given on one of the AX to FX leads, a flip-flop in the switch memory is set which, in turn, puts out a signal on the corresponding K-lead. This calls for the operation of the transmitting switch that bridges together channel X and the regular channel requesting the switch. Simultaneously, a voltage change is made on the CSX lead, which causes the IF carrier supply for channel X to be turned off. The assignment signal also flows to the encoder and memory circuit, where it is translated into one of six possible two-out-of-four signal combinations for presentation to the tone transmitter of the order channel by way of output leads C1 through C4. The actual presentation is not made immediately, however. It is deferred for a short period of time pending (a) receipt of a signal from the transmitting switch verifying the fact that the switch ordered to operate has actually operated, and (b) the successful completion of a check, termed the "inhibit-timing check", which indicates that the transmission failure or degradation has actually occurred in the switching section under consideration rather than in the preceding switching section. Switch verification signals are received by the switch-verification circuit over leads P1 to P6 shortly after the orders are given over leads K1 to K6. The output of the verification circuit is a signal to the valid switch order circuit. This, if and when accompanied by an appropriate signal from the inhibit check circuit, allows the receiving switch order to proceed.

The third usage of the assignment signal is by the channel-selected circuit, which produces an output signal on the SX lead as soon as any of the six assignments possible with channel X have been made. The SX signal starts the operation of the inhibit-timing circuit, which immediately makes a voltage change on the ITX lead lasting for 8 milli-

seconds before reverting to its normal level. This ITX signal, in turn, (a) actuates the service-request circuit, (b) inhibits the action of the valid switch order circuit and (c) enables the inhibit-memory circuit. The service-request circuit, through the SR lead, gates the channel-status circuits so that no changes in their output signals can be made during the inhibit-timing interval even if changes actually take place on the status channels themselves.

If the transmission fade or failure that caused the channel-status indication to change from good to bad actually exists in the switching section under consideration, the inhibit-timing interval runs its course with no other event until the end of the interval. At this point, provided the verification signal from the transmitting switch has been received, the valid switch order circuit is permitted to gate the encoder and memory circuit so that flip-flops associated with the four outgoing order leads are set in accordance with the code already established. This action constitutes an order to the signaling equipment and thus to the distant receiving logic.

If the transmission failure is in the switching section preceding the one under consideration, the act of bridging a protection channel onto the regular channel does not restore transmission. On the contrary, the status of protection channel X changes from good to bad as soon as the X channel IF carrier supply stops and the monitor and status-detector circuits respond to the change. This occurrence, if it takes place, is during the inhibit-timing interval and causes a voltage change on the channel X status lead, designated BS7. Such change within this particular time zone is interpreted as a failure in the preceding switching section. The BS7 lead connects, in addition to other points, to the inhibit-memory circuit. An interruption in its normal potential during the ITX enabling period produces (a) a 110-millisecond pulse on the IAX to IFX output lead corresponding to the regular channel that reported the transmission failure and (b) a short pulse on the A1 lead going back to the inhibit timing circuit. This latter signal sets a flip-flop which prevents any further action towards carrying out the switch order. The pulse on the IAX to IFX lead serves to make the channel-status indication good again for a period of time ordinarily ample for the previous section to make good its own failure. The consequences of the channel being given a good indication are to release the assignment and to restore the SX lead to normal. This action, together with the presence of a signal on the IX lead, causes the memory reset circuit to restore the transmitting switch memory to normal and thus to release the previously operated transmitting switch. The inhibit-memory circuit was informed of the identity

of the failed channel by a signal on one of the RAX to RFX leads which was produced by the transmitting switch memory when it first received the assignment information.

In the instance where the transmission failure is found to be in the switching section under consideration and the receiving switch order is allowed to proceed at the end of the inhibit-timing interval, the next work of the transmitting logic is to check that the receiving switch order is acted on. Arrangements are provided in the distant receiving switch logic and control circuits to interrupt, momentarily, the transmission of tone on the status channel corresponding to the protection channel being used, after the operation of the receiving switch has been completed. Such a momentary interruption received on the BS7 lead within what is termed the "switch-initiation interval" constitutes a signal that the receiving switch has been effected. This timing interval runs for about 65 milliseconds and is started by a signal on the SIX lead through the action of the inhibit-timing circuit at the end of its 8-millisecond interval and coincident with the transmission of the receiving switch order code. Throughout the switch-initiation timing interval, the SR lead, through the action of the service-request circuit, again prevents any change being made in the status indications of the channel circuits.

After the timer runs its course, a signal on the SITX lead causes the service-request circuit to revert to normal. This completes the normal switch operating cycle. The channel-status circuits are now free to respond to other changes that may take place.

In the event the switch-initiation check signal is not received, the switch-initiation timing circuit generates a short pulse on the SFX lead at the end of the 65-millisecond interval. This signifies that the initial attempt to switch failed and that another attempt should be made using the other protection channel, if available. The SFX lead pulse brings this about by causing the channel X status circuit to mark itself bad for a period of approximately 140 milliseconds. While this is not the true condition of the channel, such a signal has the effect of forcing a re-assignment to the other protection channel. At this point the transmitting switch memory is reset in preparation for the new assignment and the new switch attempt.

When a fade on a channel passes and the status indication on the BS1 lead, for example, turns "good", both receiving and transmitting switches are returned to normal as soon as possible to free the protection channel for other assignments. Assuming no other request is being served at the time, a good indication on BS1 will immediately set the memory flip-flop in the channel-status circuit so as to provide a good indication

on the GA lead. This causes the assignment then in effect, AX for example, to be withdrawn. This does not cause any immediate change in the memory or controls associated with the transmitting switches, but it does cause immediate reversion to normal of the receiving switch order code. This constitutes a new order to the distant receiving logic signifying that the receiving switch in the X channel group, then operated, should now be released and a check signal sent back to the transmitting logic upon completion of the release. The withdrawal of the AX assignment also, at this time, causes the SX lead to revert to normal. This, together with the condition then existing on the CSX lead, causes a timer to start in the release fail timing circuit. The purpose of this is to check that the receiving switch release signal occurs within the time period allowed, nominally 55 milliseconds. The check signal, as during switch operation, is in the form of a momentary interruption of the voltage normally present on the BS7 lead. When this signal occurs, the SX lead being normal, the memory reset responds and restores to normal the transmitting switch memory circuit, which in turn releases the transmitting switch and restores the normal potential to the CSX lead. The latter, in addition to causing the IF carrier to be restored to the channel X transmission circuit, also signals the release fail timing circuit that the switch release operation has been completed. Both the timer and the service request circuit restore to normal.

In case trouble should develop and delay or prevent the transmission of the switch release signal, the 55-millisecond timer will run its course, set a flip-flop and produce a release fail signal on the RFTX lead. This results in an alarm, locked-in until released by maintenance personnel. During this period the transmitting switch is kept operated, the regular channel is marked good, and the protection channel marked bad. This is the best policy to follow according to the information available to the transmitting logic under these circumstances.

6.2 *Organization of Receiving Logic*

The receiving logic associated with one protection channel is shown in block diagram form in Fig. 15. The second protection channel requires a duplication of this circuitry.

When transmission is good on all regular channels, protection channels are not required, and a "no switch required" signaling code exists on the R1 to R4 switch-order leads incoming to the decoder circuit. An order for the protection channel to be switched in as a substitute for one of the six regular channels consists of the establishment of the two-out-of-four code combination corresponding to the channel affected. Any case of

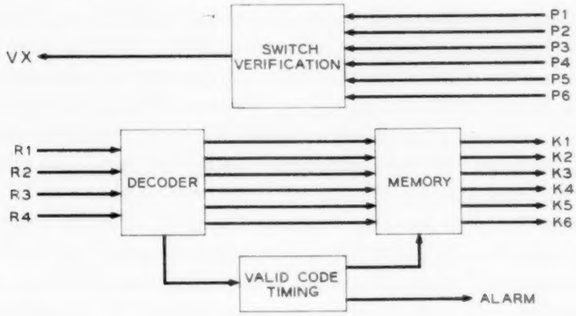

Fig. 15 — Block diagram of receiving logic.

one or three leads energized, or none at all, constitutes an invalid code which, if it persists for a sufficient period of time, will cause an invalid-code alarm to be given.

A two-out-of-four code is translated to a one-out-of-six code for presentation to the receiving switch memory circuit. The decoder acts instantly, but registration in the memory is deferred for about five milliseconds to make sure the switch request is valid and is not the result of impulse noise.

During the transition from the normal four-signal code to a two-signal code or vice versa, there will almost always be a momentary three-signal code indication as a result of differences between order tone signals in respect to transmission and detection times. This short, transient invalid-code indication normally serves to advise the valid-code timing circuit that a change in code has taken place and that a new timing cycle should be started. At the end of the timing period, an enabling signal is given to the switch memory circuit permitting the one-out-of-six input signal then present to operate its corresponding flip-flop memory and, through this, to cause the operation of the designated receiving switch. The enabling signal lasts just long enough to insure operation of the flip-flop. This strengthens the protection of the memory against unwanted interference.

Against the possibility that the change in code might be so perfect that no invalid-code transient would be available for triggering the timer, the latter is provided with a self-recycling circuit so that every six milliseconds an enabling pulse is generated, giving the memory an opportunity to readjust itself to any change in code, unaccompanied by an invalid-code transient that may have taken place in the preceding interval.

The operation of the receiving switch, or its restoral to normal, causes a voltage change on the appropriate lead in the P1 to P6 group, which causes the switch-verification circuit, to produce a four-millisecond pulse on the VX or VY lead, depending on the protection channel involved. This in turn causes a momentary interruption of tone on the corresponding status channel which, as previously explained, acts as a verification signal to the transmitting logic.

The preceding two sections describe the most general action of the switching logic. In some situations, elements can be omitted. As examples, with a transmitting baseband switch, the inhibit-timing check is omitted, and with dc reporting, the trnsmitting encoder and the receiving decoder are replaced by direct wire interconnections.

6.3 *Logic Circuit Elements*

The direct-coupled, transistor-resistor logic (TRL) gate is the basic building block used in the combinational and sequential operations of the control logic. This type of gate circuit, implemented by the silicon diffused-base NPN transistor (WE 16A), offers pronounced advantages in respect to over-all reliability, circuit simplicity and flexibility. Its speed of signal propagation, in the microsecond range, is ample for the needs of protection switching logic. The circuit and corresponding logic symbol are shown in Fig. 16.

If the potential at the input of any of the gating resistors, R_g, is raised to the prescribed level, the transistor conducts heavily (saturates), and the output signal voltage is nearly zero or ground potential. With all inputs at nearly ground potential, virtually no current flows in the collector electrode, and the output voltage then is relatively positive. The more positive level of input or output voltage is considered, for purposes of logical analysis, as state "1". The zero or less positive level of voltage is considered as state "0". Thus, with all inputs at state "0",

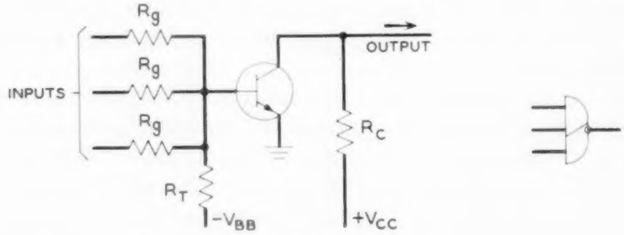

Fig. 16 — Basic TRL gate and symbol.

the output is at state "1" and, with one or more inputs at state "1," the output is at state "0". This is the action of an OR gate with the output signal inverted. The circuit is usually described as a NOR gate. In the symbol the slanting line signifies the OR function, while the small circle signifies inversion in the sense of the output signal. A gate of this type with only a single input is described simply as an inverter. In the symbol for this the slanting line is omitted.

The gate design adopted is conservative and allows for as much as 350 millivolts of noise at the base of the transistor. It will accommodate as many as four inputs and will drive, simultaneously, as many as four output stages of the same design or other circuits of equivalent loading effect. Interstage connections between the transistors of two gates are shown in Fig. 17. In some cases six- or seven-input gates are required, and two transistors are used with their collectors connected in parallel to a common resistor.

The limited amount of memory required in the logic dictates the use of a conventional, saturating type of bistable circuit, or flip-flop, for each bit of memory. The circuit and its representation are shown in Fig. 18. The control of certain memory functions is facilitated through the use of a special flip-flop design in which setting and re-setting are both effected over the same lead. This circuit is always used in conjunction with a two-transistor, bi-lateral switch that allows the flip-flop set-reset lead to be opened or closed as required. The combination of circuits and the symbols used to represent them are shown in Fig. 19.

The principal feature of the assignment circuit is another modified form of bi-stable circuit, similar to a flip-flop but differing in the manner of its control. This circuit element, together with two diode-type AND gates used to enable or inhibit its action, are shown in Fig. 20. The pur-

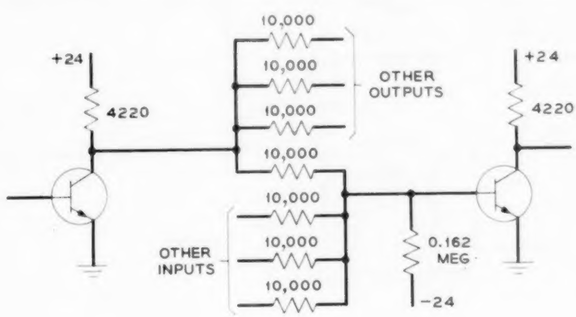

Fig. 17 — The TRL interstage connections.

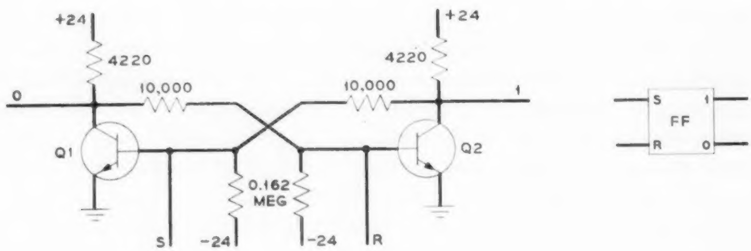

Fig. 18 — Standard flip-flop circuit and representation.

pose of this circuit is to assign one of the two protection channels to the regular channel corresponding to the bi-stable circuit in question and to insure that this protection channel cannot be assigned to a second regular channel at the same time. Preference as to which protection channel should be assigned to a given regular channel under normal conditions is determined by the position of the switch associated with the 0.1 mf capacitor.

If the normally preferred protection channel is not available for any reason, a small negative potential is present on one of the six AND gate leads associated with it. This allows forward current to flow through the diode and thus causes a slight negative bias on the base of the transistor, preventing the latter from operating. In this situation, the other "second-choice" transistor will operate provided, of course, its

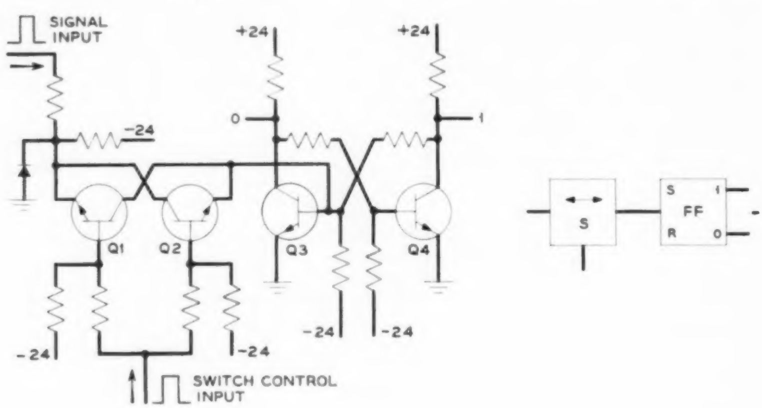

Fig. 19 — Bi-lateral switch and special flip-flop with symbols.

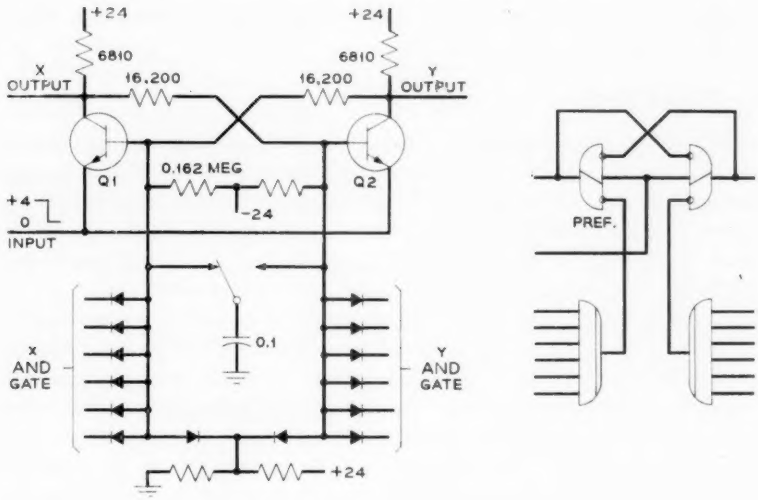

Fig. 20 — Assignment flip-flop with diode AND gates and symbols.

associated protection channel is available as signified by no current flow in any of the AND gate diodes.

In the symbolic representation of this circuit feature, the vertical line through each of the lower gates signifies the AND function. Each transistor of the bi-stable circuit, in terms of its logic function, may be represented as an OR gate (diagonal line) with respect to a "1" output in which two of the inputs are inverted as shown by the small circles.

Fig. 21 shows the circuit of a monopulser typifying several that are used in the logic for establishing time intervals for certain functions.

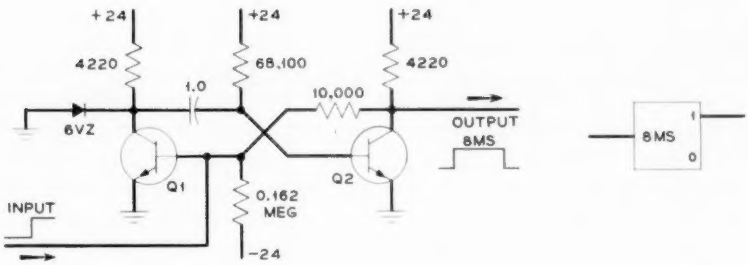

Fig. 21 — Typical monopulser and symbol.

Another important type of timing circuit employed is the delay timer shown in Fig. 22.

6.4 *Logic Diagrams*

Fig. 23 is a logic diagram of the regular and protection channel-status circuits and the channel-available and service-request circuits, indicated as blocks in Fig. 14. In addition to the TRL gates for handling the various input signals, this diagram shows the status-memory flip-flops, the bi-lateral switches and the timers employed. Most of the time the memory flip-flops and, consequently, the status indications on the GA and GX leads follow directly any changes in indications on the BS1 and BS7 leads. If the logic is busy making or releasing a channel switch, or if there is no protection channel available, signals generated in the service-request or channel-available circuits cause the bi-lateral switches to be opened and thus prevent any changes in the positions of the flip-flops until such periods have ended. The 50-microsecond delay timer in the control path of the regular channel switch serves to prevent unwanted operation of the switch during a small gap between the end of the inhibit-timing interval (signal on ITX lead) and the beginning of the switch-initiation timing interval (signal on SITX lead). The 5-millisecond delay circuit in the path of the protection channel status signal (BS7) lead prevents improper operation in the event of a general failure of the status-signaling transmission facility. The delay allows the logic of the alarm circuit (not shown) to determine that all channels have failed and to place switch-inhibiting signals on the SR and NSR leads.

A manual switch of a regular channel may be made by grounding the MSA lead. A positive voltage on the LOA lead (via a lockout control key) will prevent an automatic switch by the regular channel and, on the

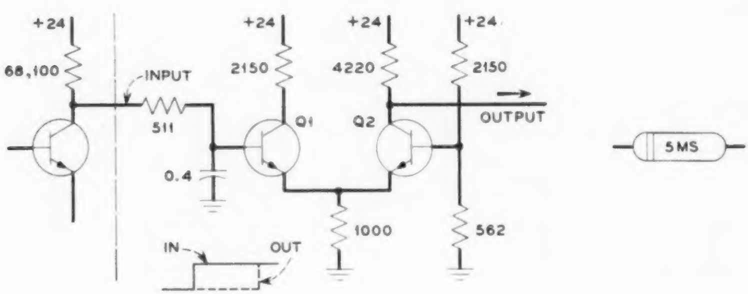

Fig. 22 — Delay timer and symbol with typical input circuit.

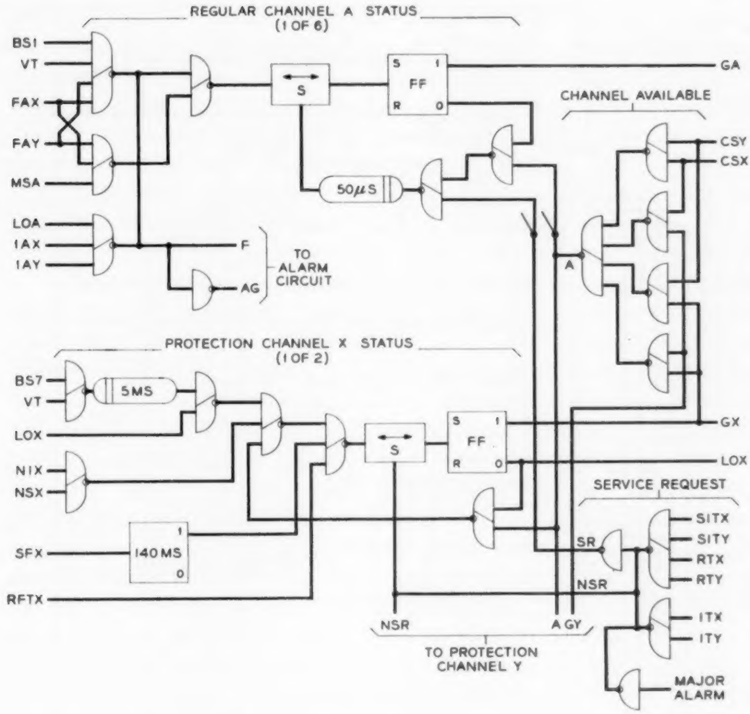

Fig. 23 — Logic diagram of regular and protection channel-status circuits, with associated channel-available and service-request circuits.

LOX lead, by the protection channel. The VT lead is used to inhibit any failed channel indications that may be received when it is desired to make out-of-service operational checks of the logic.

Fig. 24 is a logic diagram of the assignment circuit, shown in detail for three regular channels. Signal leads from the regular channel-status circuits enter from the left, those from the protection channels from below. The output leads, signifying by their designations the various possible assignments, are on the right. With the system normal, all input leads have a positive voltage and all output leads zero voltage.

The assignment flip-flops, discussed in connection with Fig. 20, are in the middle of the diagram. Each flip-flop output becomes one of two inputs to a TRL gate whose output, in turn, when it becomes positive, constitutes an assignment signal. In addition, these latter outputs, after

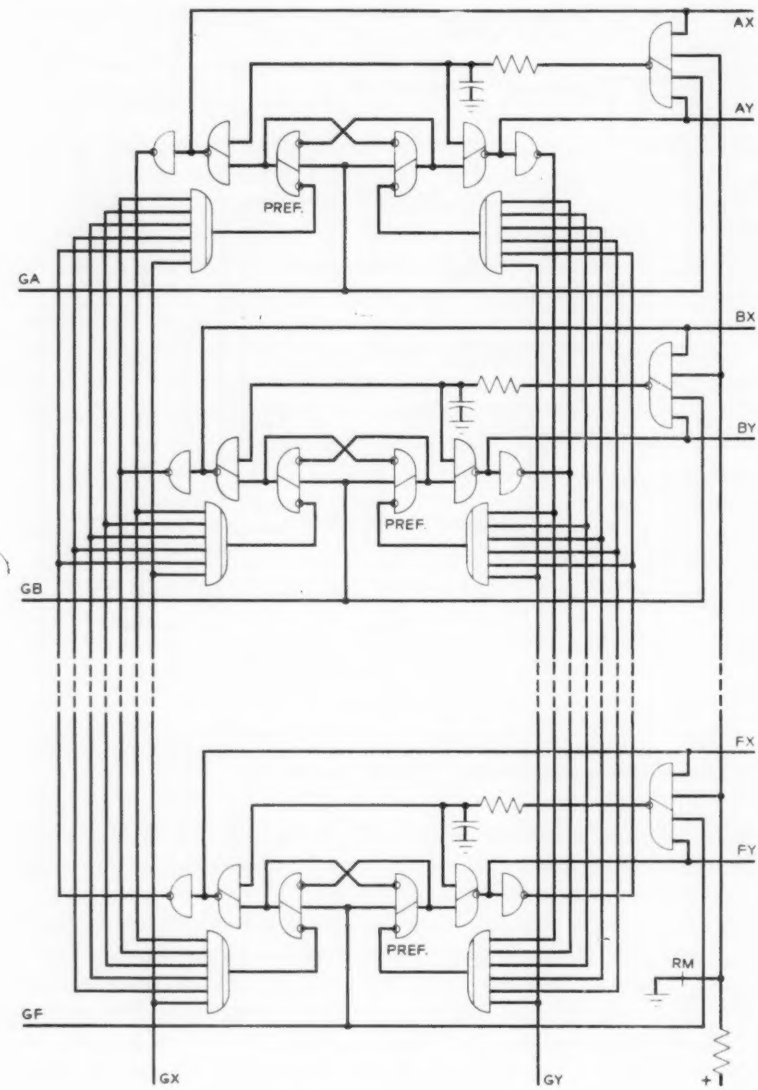

Fig. 24 — Logic diagram of the assignment circuit.

inversion, actuate all AND gates associated with the chosen protection channel, excepting that associated with the regular channel requesting the assignment. This prevents the chosen protection channel from being assigned to any other regular channel that may request a switch. Subsequent bids will result in the assignment of the second protection channel, regardless of the normal preference.

If an attempt to complete a switch is unsuccessful, and then the ensuing attempt on the other protection channel is also unsuccessful, it is essential that no repetition of this cycle be allowed until the difficulty is corrected. If this situation arises, both the GX and GY leads are at zero potential temporarily, along with a zero potential on the channel lead. Since no assignment signal (i.e., no positive voltge on output -X or -Y lead) exists under this condition, the gate at the right of the channel circuit is cut off and, as a consequence of its positive output signal, is locked in this position until either the regular channel withdraws its bid for a switch (GA lead again becomes positive), or until the RM key is manually operated by maintenance personnel. During such a lockout period the protection channels are available for assignments to the other regular channels. Although the lockout gate is momentarily cut off each time its channel input lead goes to zero, the effect of this is delayed by the RC circuit in its output until the assignment signal is established, whereupon the gate is restored to its normally operated condition.

The memory functions associated with the transmitting switch control and with the order-tone transmitter for protection channel X (as an example) are shown in Fig. 25. At the top of the figure are two of the six switch memory flip-flops. An assignment signal (positive voltage) appearing on one of the six leads, AX to FX, "sets" the corresponding flip-flop, and thus establishes a positive voltage on the K-lead, which in turn causes the operation of the proper transmitting switch. The same memory information in inverted form is also available on one of the RAX to RFX leads. This is utilized in certain circumstances by the timing and checking circuits. The K-lead voltage also operates one of the TRL gates connected to the CSX lead, reducing the latter's normally positive voltage to zero. This causes disconnection of the carrier supply and also notifies points within the logic of channel X's busy condition.

The switch-verification logic checks for agreement between the signals given a transmitting switch to operate or release and the signals received via the P-lead showing that the switch responded as directed. The output voltage of this circuit is normally zero, but becomes positive during the transient operating or releasing interval of the switch or in case of trouble when no verification signal is received.

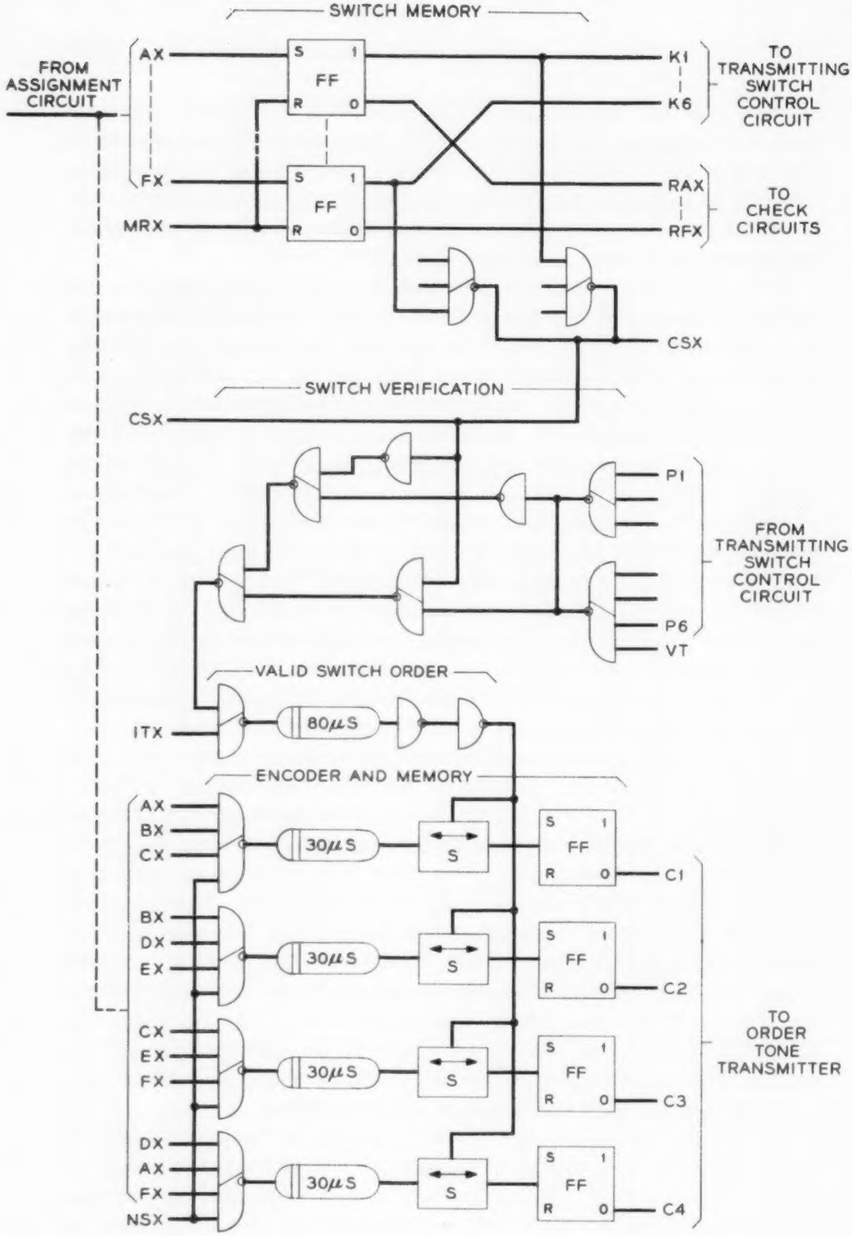

Fig. 25 — Logic diagram of switch memory, order encoder and related features.

The valid switch order circuit, under joint control of the transmitting switch-verification signal and the inhibit-timing signal, determines when it is time to permit a change in the setting of the memory flip-flops associated with the encoder and thus present the tone transmitter with a switch order. With the protection channel not in use, all four order-tone flip-flops are in the reset state with positive voltages on leads C1 through C4. The four TRL gates at the lower left comprise the encoder. Normally, the gate outputs are at zero voltage level owing to a positive voltage on the NSX lead. When a given assignment signal is established, the inputs of two of the encoder gates are made positive, and a microsecond later, due to action in the timing and checking circuit, the NSX lead voltage drops to zero. This combined action allows the outputs of two of the four TRL gate outputs to become positive, which constitutes basic encoding operation. The two positive outputs are delayed about 30 microseconds in transmission to allow time for the establishment of a positive voltage on the ITX lead, the release of the normally operated 80-microsecond timer, and the opening of the bilateral switches. At the end of the ITX signal, provided the switch-verification signal is present, and after the delay involved in operating the 80-microsecond timer, the bilateral switches close and thus permit the two flip-flops to be set in accordance with the encoder output signals. The 80-microsecond delay is required to allow adequate time for changing the assignment and its encoded equivalent in the event of failure to complete a switch on the first attempt and also to prevent premature setting of the order flip-flops in case the inhibit-timing interval check indicates the transmission failure to be in the previous switching section.

Fig. 26 shows the logic of the timing and checking circuits for channel X. The channel-selected circuit (lower right) receives the assignment signal simultaneously with the switch-memory and order-tone encoder. The output of the channel-selected gates is a positive voltage on the SX lead and a zero voltage on its inverted counterpart, the NSX. The change in SX lead voltage resets the inhibit flip-flop and triggers the inhibit-timing monopulser. The latter makes the ITX lead positive for a period of 8 milliseconds. The inhibit-memory circuit (left middle) contains the logic and temporary memories for producing a positive signal of 110-millisecond duration on the IAX to IFX lead corresponding to the channel assignment, in the event the voltage on the protection channel status lead, BS7, becomes zero during the ITX interval. This, if and when it occurs, indicates that the trouble is in the preceding switching section. The IAX to IFX lead signal forces the regular channel-status circuit to give a good indication long enough for the preceding section to complete a switch and restore transmission. The operation of one of the 110-milli-

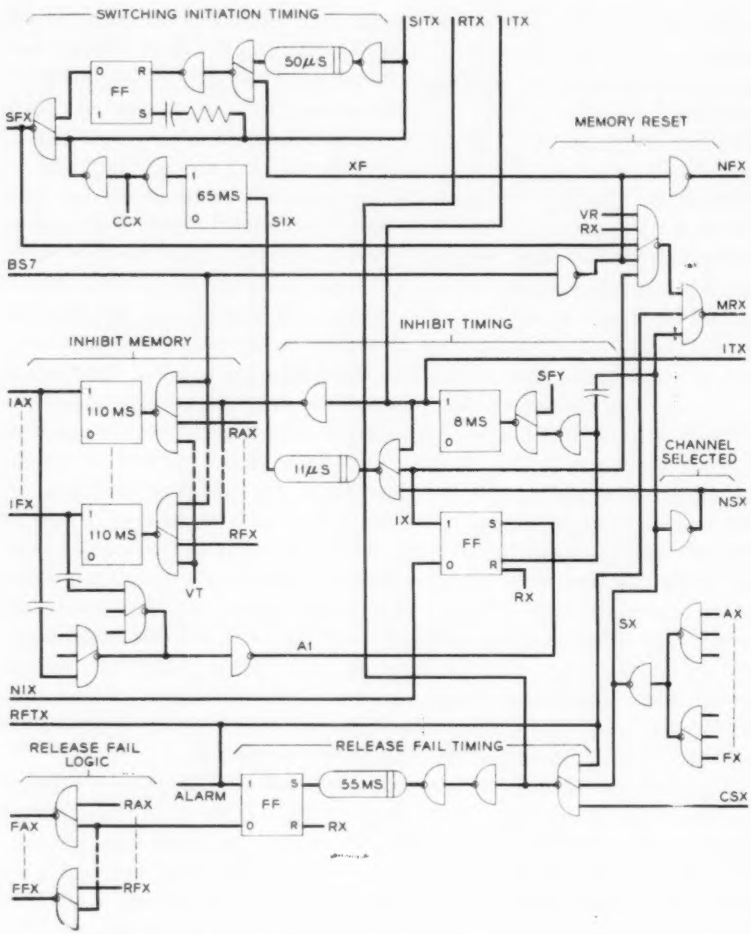

Fig. 26 — Logic diagram of timing and checking circuits.

second monopulsers also generates a transient impulse that sets the inhibit flip-flop which, in turn, drives the IX lead positive. This action prevents the operation of the inhibit-timing output gate and thus prevents the start of the switch-initiation timer, which otherwise would follow the end of the 8-millisecond ITX interval. At the end of this latter interval, the service-request circuit permits the channel-status circuit to respond to the 110-millisecond inhibit signal, the GA lead (Fig. 23)

becomes positive, the assignment signal is withdrawn, and the SX lead potential reverts to zero. This, coupled with a positive IX lead, actuates the memory-reset circuit to make the MRX lead positive and thus resets the switch memory. The NSX lead, in reverting to its positive voltage state, normalizes the encoder without the order having been transmitted.

If no loss of positive voltage occurs on the BS7 lead during the 8-millisecond interval, the regular switching cycle is allowed to continue. The return of the ITX lead to zero potential, acting through the valid switch-order circuit, allows the encoded order to be transmitted and, coincidentally, causes the inhibit-timing gate output to become positive. This, after an 11-microsecond delay (to avoid response to short transient signals), produces a positive voltage on the SIX lead, which marks the beginning of the switch-initiation timing interval, during which the logic looks for an indication that the desired receiving end switch has been made. The SIX signal triggers the 65-millisecond monopulser, which drives and maintains the SITX lead positive for this period. The switch initiation flip-flop is set at this time for later use. The SFX lead potential, normally zero, is held at this level by the positive SITX lead input to the output gate. In the normal case, during this period, the BS7 lead will lose its positive potential for a few milliseconds as a signal from the receiving end logic that the receiving end switch has been completed. This occurrence will cause the XF lead to become positive and, by way of the gate leading to the switch-initiation flip-flop, cause the latter to reset and produce a positive voltage on its output lead. This will continue to hold the SFX lead potential at zero after the 65-millisecond period of positive potential on the SITX lead ends. In the absence of a signal on the SFX lead, no further action on the part of the X channel logic is required at this time.

In case a momentary interruption of the positive potential on the BS7 lead fails to occur before the 65-millisecond timing interval ends, the flip-flop will not have been reset before the SITX lead potential returns to zero. This permits the SFX lead potential to become positive. The same action, however, starts the delay timer which, after 50 microseconds, resets the flip-flop, and this in turn causes the SFX lead potential to revert to zero. The 50-microsecond pulse thus created on the SFX lead triggers the 140-millisecond monopulser in the channel X status circuit, which results in a second attempt to complete a switch, this time using channel Y, if available.

The circuit in the lower part of Fig. 26 checks that the release of a protection channel, when called for, takes place within the allotted

interval of time. A regular channel becoming good causes the assignment signal to be withdrawn. This changes the voltage of the SX lead to zero and that of the NSX lead to positive. The latter causes the encoder and memory circuit to revert to normal, which constitutes a signal to the receiving logic that the receiving switch previously operated should now be released. When the SX and CSX leads are both at zero voltage, the lower NOR gate acts to produce a positive signal on the RTX lead, which starts the 55-millisecond delay timer. Normally, before this timer runs its course, there is a momentary interruption of the positive voltage on the BS7 lead, signifying that the receiving switch has released as ordered. This, in combination with zero voltage on the SX lead, actuates the memory-reset circuit to put positive voltage on the MRX lead, which resets the switch memory flip-flop. This, in turn, releases the transmitting switch. The CSX lead voltage becomes positive, which resets the 55-millisecond timer and thus forestalls the setting of the release-fail flip-flop. As soon as the release-verification signal ends (BS7 lead potential becoming positive), the entire protection channel becomes normal and available for a new assignment.

If the release-verification signal should not be received before the 55-millisecond timer runs out, the release-fail flip-flop will be set which, by making the RFTX lead positive, will hold the transmitting switch operated, mark the protection channel bad in its status circuit, and present a release-fail signal to the alarm circuit. In conjunction with the signal present on the affected lead in the RAX to RFX group, the flip-flop also causes a positive voltage on the FAX to FFX lead, which marks the regular channel bad in its status circuit. These actions prevent any further change in the regular or protection channel logic until maintenance personnel have determined that the circuits should be normalized and the alarm retired. This is done by a key operation which applies positive voltage to the RX lead for resetting the flip-flop.

The logic details of the receiving circuit for one protection channel are shown in Fig. 27. The decoder is on the left. With all order-tone channels normal, there is a positive voltage on all four input leads and zero potential on all six decoder output leads. The latter connect to the bilateral switches of the memory and also to a seven-input, two-section gate for indicating the presence of a valid code. The four-input gate in the lower part of the decoder has a positive output when all order-tone channels are normal. This represents the valid "no switch required" code. When a two-out-of-four code is established, the voltage of this gate returns to zero, while that of one of the six channel gates becomes positive. Except during transitions, there is a positive voltage on one of

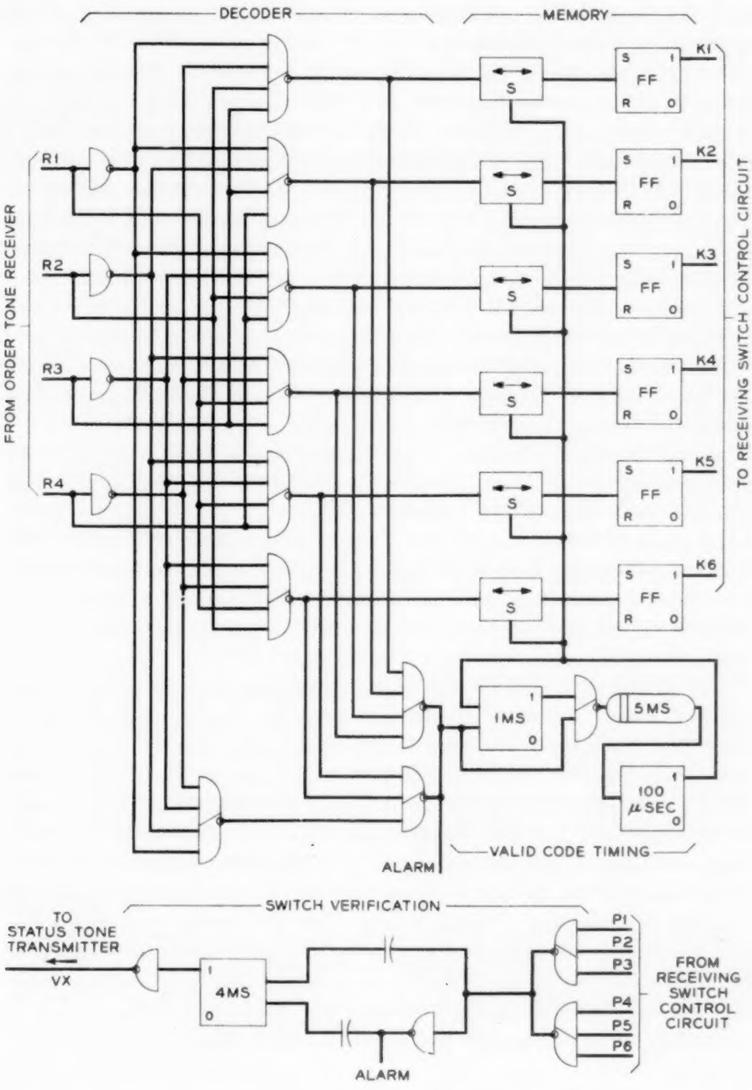

Fig. 27 — Logic diagram of receiving-circuit decoder, memory, checking and timing features.

the seven input leads and thus zero voltage on the output lead of the valid code gate at all times.

During a change in the incoming code, a transient one-out-of-four or three-out-of-four combination is a normal expectancy. Either will produce a momentary positive voltage on the output of the valid code check gate, which triggers the one-millisecond monopulser of the timing circuit. This resets the five-millisecond delay timer. At the end of the monopulser output or at the end of the transient invalid code indication, whichever occurs last, the five-millisecond delay timer starts to operate. On completion of its operation, it triggers the 100-microsecond monopulser, which operates the bilateral switches long enough to allow the operation of the channel memory flip-flop corresponding to the decoder output gate having positive voltage at this time. This arragement of delay and limited access to the flip-flops affords a high degree of protection against interference and simulated orders due to impulse noise.

In case the usual transient triggering impulse is not developed when a code change takes place, reliance is placed on the feedback circuit between the output of the 100-microsecond monopulser and the input of the one-millisecond monopulser. This keeps the three timing elements constantly recycling with a periodicity of about six milliseconds. While this does not guarantee the full five-millisecond time lag normally provided, it insures that a change in the status of the decoder does not go unrecognized for more than six milliseconds.

The setting of a memory flip-flop applies positive voltage to the associated K-lead which, through the receiving switch control circuit, causes the designated receiving switch to operate. Shortly after this, a verification signal in the form of a positive voltage on the corresponding P-lead is received which drives the output of the verification gate to zero. By the way of the inverter, this becomes a positive-going transient which triggers the four-millisecond monopulser. During the verification interval, the potential of the VX lead becomes zero and the status tone associated with the X protection channel ceases. This tone break constitutes the receiving switch-verification signal utilized by the transmitting logic. On the release of a receiving switch, the P-lead potential reverts to zero. This causes a positive-going transient via the upper capacitor, which again triggers the monopulser and creates an interrupted tone-verification signal as before.

VII. SYSTEM MAINTENANCE AND EQUIPMENT ARRANGEMENTS

Most active equipment will be tested periodically on a routine basis. The circuits to be maintained are easily removable for insertion in a special test set. Ref. 5 comprises a description of this set, together with

the principal features of the testing procedure. Before a circuit is removed, the proper manual switching operation is made either to isolate the circuit from the system or to bring it into a passive state which allows its removal. The function of active circuits is replaced, either automatically or manually, before such circuits are completely removed. A large number of cards in the transmitting and receiving logic, however, cannot be removed for testing without first disabling the automatic portion of the switching system by an override status quo action. No preventive maintenance is anticipated on such cards.

If a trouble occurs somewhere in the switching system, alarms normally are issued. The type of alarm and the indication of lamps permit a fairly quick localization of the circuit area in trouble. Pin jacks located on the front plate of certain circuits are used to pinpoint the failed circuit. A failure in the transmitting logic is more difficult to localize. A special logic-tester is used in this case. Furthermore, a total of 12 message registers located in the transmitting logic count the number of completed switch operations. If an excessive number of switch operations are registered for a particular channel, trouble must be suspected, and the necessary actions can be taken to eliminate this condition.

Companion articles describe in some detail the equipment features[6] and the power supply[7] for the protection switching system. Most circuits for the switching system are built on cards which are easily inserted and removed from the equipment bays. Four types of 11-foot bays are used: the IF switching bay, the baseband switching bay, the control bay for tone reporting and the control bay for dc reporting.

The switching system is powered by a very reliable plant delivering +24 volts and −24 volts. A high degree of reliability and a stable voltage even under emergency conditions are achieved by the use of buffer batteries and the firm ac power equipment of TH as a source. The power plant is mounted in separate bays.

VIII. ACKNOWLEDGMENTS

Many members of the Laboratories have contributed to the development of the automatic protection switching system. Among them the authors wish to mention J. J. Degan, W. R. McClelland, R. K. Townley, H. I. Maunsell and R. H. Higgins.

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Test Equipment for the TH Radio System

By E. W. HOUGHTON and E. J. DRAZY

(Manuscript received June 12, 1961)

To maintain the high performance objectives of the TH radio system under operating conditions requires the regular use of accurate field test equipment. The necessary test instruments are incorporated into mobile test consoles, one for each major subdivision of the TH system. Three of these are described in some detail in this article: 1) the broadband radio transmitter-receiver test set; 2) the FM terminal test set; and 3) the protection switching test set.

I. INTRODUCTION

It became apparent early in the development program for the TH radio system that to operate the system in the field and to attain its high performance objectives would require the regular use of accurate field test equipment. Therefore, the planning and development of field maintenance facilities was done concurrently and cooperatively with, and continued throughout, the development of the rest of the TH system.

At the outset it was decided to include in a mobile console all of the test instruments required to make all routine and some trouble location tests on each major subdivision of the system. From this philosophy, there results a number of mobile test consoles, each named for that portion of the system which it is primarily designed to maintain. Transmitter-receiver test sets, available in all repeater stations and maintenance centers, are designed to assure optimum performance of the broadband radio transmitter and receiver and the RF and IF portions of the auxiliary channel.[1] FM terminal test sets, available wherever there are FM terminals and in maintenance centers, are designed to adjust and maintain FM terminal transmitters and receivers.[2] Protection switching test sets, appearing at all switching stations, maintain the protection switching system.[3] Currently under development is an auxiliary channel test set for testing baseband portions of the auxiliary channel.

Most of this paper is devoted to a description of the first three sets. Brief descriptions are given of maintenance center testing facilities, designed to permit trouble-shooting, repair and adjustment of units that have been removed from the system; and test techniques for straight-away video transmission and IF delay distortion measurements on interconnected links of the system.

II. REQUIREMENTS AND OBJECTIVES

Requirements on the performance of the TH system dictate what quantities are to be measured, the frequency bands, and the measurement accuracies. Ideally, test set errors should be very small on an absolute basis. Practically, they must only be small compared to expected instabilities in the quantities being measured. This usually significant gap between possible and practical accuracies in many cases results in large dividends in measuring convenience, compactness and reliability of the test equipment.

2.1 *Test Signals*

Just as the system employs analog transmission signals in three distinct frequency bands to accomplish its purposes, so are maintenance instruments required for each of these bands. Signals appear in a 60-cps to 10-mc baseband, a 58-mc to 90-mc intermediate-frequency band and a 5925-mc to 6425-mc radio-frequency band. All of the signals measured with the first two sets are analog in character. Digital control signals (dc and pulses) are measured in the protection switching test set.

2.2 *Measured Quantities*

A wide variety of quantities must be measured with the transmitter-receiver and FM terminal test sets: impedance, amplitude transmission, frequency, frequency deviation, power, etc. Altogether, there are eleven different quantities, some of them in each of the three bands. To simplify operating instructions, similar quantities are measured by similar methods in the several different bands. This approach, together with a maximum use of common control, amplification and detection circuits, results in a considerable reduction in the equipment required to accomplish the totality of measurements.

While the variety of quantities to be measured by the protection switching test set is not so great, nearly 1000 dc and time interval test observations can be made on the 50 different units of the system.

A summary of the quantities measured in the three signal bands is

TABLE I — QUANTITIES MEASURED BY TH RADIO TEST EQUIPMENT

Quantity	Test Set*	Frequency Band†			
		DC	Video	IF	RF
Frequency	1, 2, 3		x	x	x
Noise figure	1				x
Power	1, 2, 3		x	x	x
Gain or loss (vs. Freq)	1, 2		x	x	x
Reflection coefficient (return loss) (vs. Freq)	1, 2			x	x
FM deviation	2		x	x	
FM deviation sensitivity	2		x	x	
FM receiver linearity	2			x	
FM transmitter optimum linearity	2		x	x	
Square wave response	2		x		
Time interval	3		x		
Voltage	3, 4	x	x		
Current	3, 4	x			

* 1. Transmitter-receiver test set; 2. FM terminal test set; 3. Protection switching test set; 4. DC metering test set associated with system equipment.
† Video, 60 cps to 10 mc; IF, 50–100 mc; RF, 5925–6425 mc.

given in Table I. Throughout this paper we use video (60 cps to 10 mc), IF (50-100 mc) or RF (5925–6425 mc) to designate the general frequency region in which a test set operates.

2.3 *Accuracy*

In keeping with the more exacting objectives of the TH system over its predecessor, the TD-2 system,[4] measurement accuracies are required to be higher. For example, accuracies in the 0.02 to 0.05-db class are generally required in measurements of transmission characteristics, gain, FM deviation and FM sensitivity, etc.* These are laboratory-grade accuracies, and they must be achieved on a routine basis in the field where time is not available for the patience and watchful attention ordinarily devoted to measurements in the laboratory.

2.4 *Convenience, Speed, Reliability*

Operating convenience and speed have been increased by packaging in a single console all of the instruments required for routine tests on major subdivisions of the system, by minimizing calibration adjustments, and wherever possible by using switch controls instead of patching operations to change test functions. Convenience and speed have also

* No distinction is made here between resolution and accuracy. In some cases resolution, the ability to detect changes in a quantity, is more significant than accuracy, the ability to measure absolute value of the quantity.

been increased by using frequency-sweeping techniques for transmission, impedance and linearity measurements; this also increases the accuracy and comprehensiveness of the measured result. In fact, certain system adjustments require this simultaneous presentation of performance at all frequencies in the applicable band. Finally, lost time due to test equipment error or failure has been minimized by design approaches leading to maximum calibration stability and component reliability, and by providing easy access for "maintaining the maintenance equipment."

III. TRANSMITTER-RECEIVER TEST SET

The transmitter-receiver test set is used to measure RF and IF noise figure, power, transmission, and reflection coefficient in the broadband radio transmitters and receivers and in the auxiliary channel. Sweep-frequency techniques are employed for transmission and reflection coefficient measurements in the IF band and in individual channels of the 5925–6425-mc RF band.

3.1 *Transmission and Reflection Coefficient*

The block diagram in Fig. 1 illustrates how RF sweep-frequency transmission characteristics are measured. The frequency of the RF (klystron) oscillator is automatically and continuously swept up to ± 30 mc about a central value which can be set anywhere in the 5925–6425-mc band. The output power is held constant by automatic level control. The signal level at the input of the equipment under test is controlled by the calibrated variable RF attenuator, and by connecting the circuit under test (the "unknown") to the appropriate directional coupler output. After transmission through the unknown, the signal is rectified in the microwave detector (RF detector 3) connected to the unknown output. This rectified output, displayed on the oscilloscope screen, gives a measure of transmission deviations through the unknown. A frequency-marking circuit gives a variable frequency-identification mark on the displayed characteristic. The comparing circuit (relay) puts a reference line on the oscilloscope screen during return sweep (idle) intervals. RF to IF measurements (through networks containing frequency shifters) are made by replacing the RF detector with an IF detector.

The block diagram of Fig. 2 illustrates how reflection coefficient vs. frequency characteristics are measured at RF. The microwave hybrid junction is excited by the sweeping signal generator, described above,

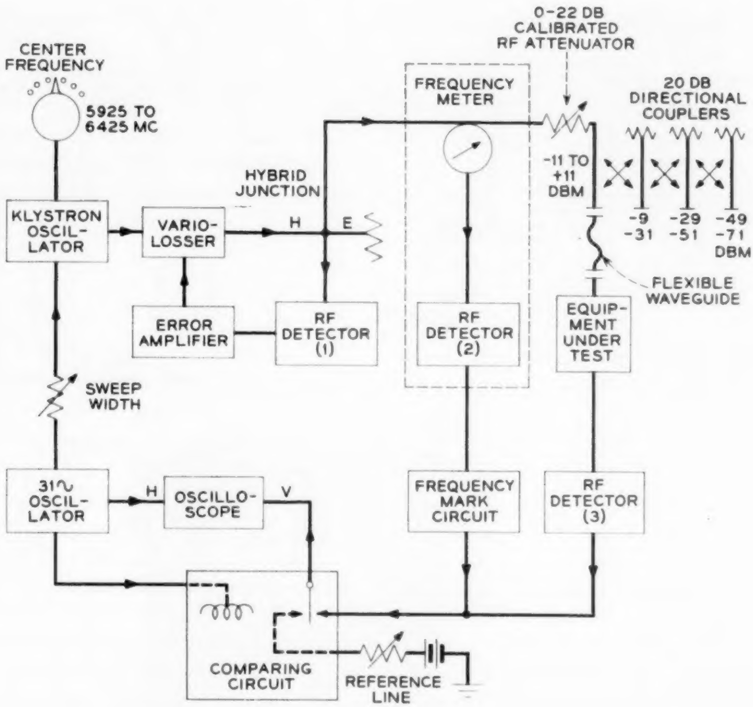

Fig. 1 — Block diagram for RF transmission measurements.

whose output level is now amplitude modulated at a 97-kc rate. Impedance mismatch of the circuit under test, connected to the hybrid junction, unbalances it. The resultant unbalance signal appears in the E-arm of the hybrid junction, where it is demodulated by the RF detector. The detector output is a 97-kc signal having an amplitude proportional to the reflection coefficient of the impedance mismatch. This 97-kc signal is amplified, rectified and displayed on the oscilloscope screen as a reflection coefficient vs. frequency characteristic. Calibration is effected by adjusting the gain of the 97-kc amplifier to give a reference line on the oscilloscope when a known impedance mismatch (reference load) is placed on the measuring arm of the hybrid junction. Subsequent adjustment of the observed characteristics to this reference line with the calibrated 97-kc attenuator gives the reflection coefficient in terms of decibels. Reflection coefficient expressed in db is called return loss; i.e., return loss $= -20 \log_{10}$ (reflection coefficient).

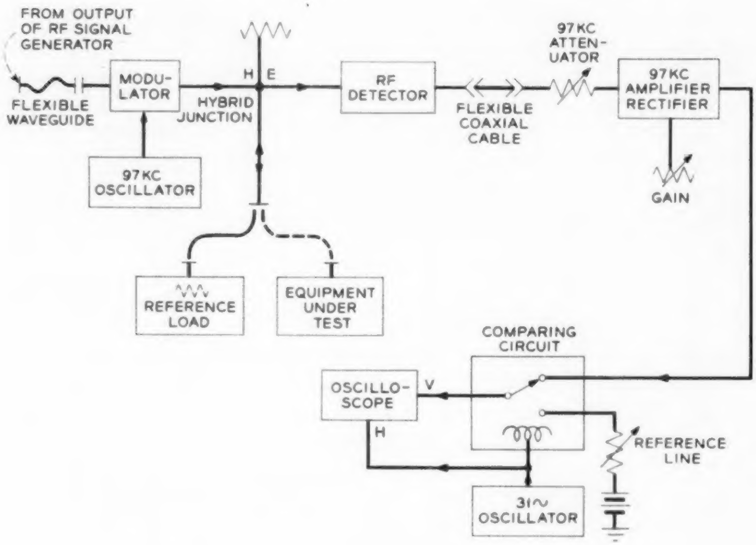

Fig. 2 — Block diagram for RF reflection coefficient measurements.

The same sweep-frequency methods are employed in making transmission and reflection coefficient measurements in the IF band. Functional block diagrams for IF sweep transmission and reflection coefficient measurements would look very similar to Figs. 1 and 2, respectively. Of course, instrumentation of these functional blocks requires different techniques and results in IF components that look quite different from their RF counterparts. However, the 31-cps time base oscillator, the oscilloscope, the 97-kc amplifier-rectifier and other low-frequency control and indicating components are used in common for IF and RF measurements.

3.2 *Noise Figure and Power*

Included in the mobile console are a noise figure test set and a power meter. These are independent of the rest of the set.

The noise figure test set measures the over-all noise figure of a radio receiver by comparing its internally generated noise with "white" noise of known magnitude. A gaseous discharge tube,[5] matched to a waveguide, furnishes the standard "white" noise. Noise figure is read on a calibrated attenuator when it is adjusted to make the sum of re-

ceiver noise and discharge tube noise 3 db higher than that of the receiver alone.

The power meter measures RF and IF power in the range of −10 dbm to +6 dbm, using a thermistor in a self-balancing ac bridge circuit. Originally designed for use with the TD-2 System,[4] the power meter is adapted for TH by using a new RF power-measuring head which is matched over the 5925–6425-mc band to the WR 159 rectangular waveguide (inside dimensions of 0.795 × 1.59 inches) used in the TH radio system.

3.3 *Physical Arrangement*

A front panel view of the transmitter-receiver test set is shown in Fig. 3. Identified on this photograph are the sweep generators, power

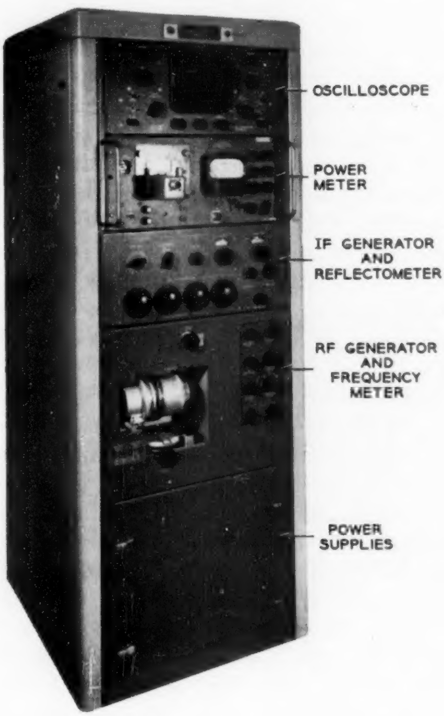

Fig. 3 — Front view of the transmitter-receiver test set.

meter and oscilloscope. The noise figure test set is located on the rear. Power supplies are located in bottom areas of the test set where exterior panel space for controls is not useful.

Connections between the test bay and equipment under test are established by 75-ohm coaxial cables for IF, and by flexible waveguide for RF measurements. Because of the relative rigidity of flexible waveguide, all waveguide connections are made at the rear of the set, leaving the front clear for ease of operation.

A considerable design emphasis was placed on ease of access and maintenance. Unitized construction, cable connectors, removable panels and a drawer-type mounting for the heavy RF generator unit allow quick access for maintenance tests, adjustments and replacement of defective units.

3.4 *Some General Design Principles*

Full realization of the speed and accuracy advantages inherent in sweep-frequency measurement methods demands that the graphic display represent only the transmission or reflection coefficient of the circuit under test and require no correction for frequency characteristics of signal source or detection components. Accordingly, major emphasis was placed on the realization of key components, such as RF and IF detectors, attenuators, hybrid junctions, resistance bridge, and reflection coefficient calibration standards, with a response constant over any 32-mc segment of the RF band and over the whole IF band. Constant response also requires that these components present, to interconnecting transmission lines, well-matched impedances over the entire IF or RF band.

Frequency characteristics due to impedance interactions over what may often be electrically long waveguide or coaxial lines connecting to the equipment under test are minimized by maintaining well-matched output impedances in the RF and IF sweep generators. Additional transmission line errors are precluded by attaching directly to the output of the equipment under test the input connector of the IF and RF detector and the measuring connector of the IF resistance bridge and the RF hybrid junction. Similarly, the RF thermistor power head is arranged for direct attachment to a waveguide output. Information is brought back to the test console through noncritical transmission lines in the form of dc or low-frequency ac.

The decision to use a sweep repetition frequency of 31 cps was made to eliminate a traditional source of error in sweep frequency measurements. This frequency provides a slight offset from multiples of the

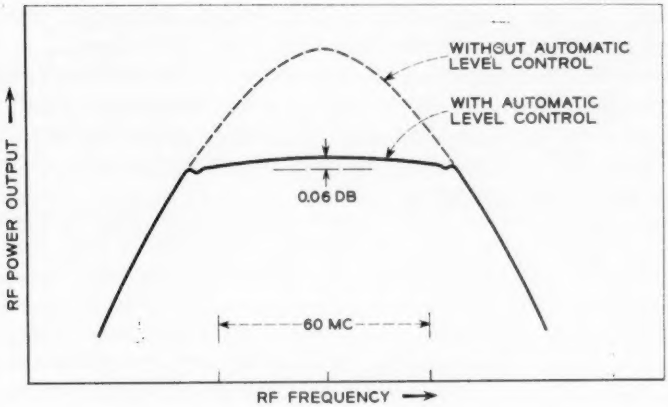

Fig. 4 — Action of the automatic level control.

power supply frequency so that any "hum" which might intrude can be immediately recognized as such and not mistakenly interpreted as a part of a transmission characteristic.* Other factors governing this selection for the sweep frequency were: avoidance of flicker on the one hand, and bandwidth requirements for the fast-acting automatic level control circuits on the other.

3.5 *RF Generator*

The RF generator has already been described briefly in Section 3.1 above and is shown functionally in Fig. 1. The RF source is a Western Electric type 449A klystron; frequency modulation is accomplished by application of a 31-cps voltage to the repeller electrode. As illustrated in Fig. 4, the amplitude modulation incidental to this process is removed by the automatic level control circuit. The loop gain of this circuit is sufficient to maintain the output constant within ± 0.03 db of the nominal $+11$ dbm throughout a ± 30-mc sweep range. The vario-losser in Fig. 1 is designed to have a smooth loss vs. exciting current characteristic. It is interesting to observe that this circuit, by regulating only the forward-traveling energy, as sampled by the hybrid junction, provides an excellent source impedance, in a manner analogous to that of a hybrid feedback amplifier.[6] The chain of directional couplers,

* Internally generated "hum" is kept to an unobjectionable level by adequate shielding, minimization of spurious ground currents, well-filtered dc power supplies and finally, by using dc heater supplies on critical electron tubes.

together with the continuously variable precision attenuator, provides output level selection over the range $+11$ dbm to -71 dbm.

When the RF detector is directly connected to the generator output, the (residual) frequency characteristic for any directional coupler and attenuator setting is less than ± 0.1 db over any 30-mc segment of the RF band; RF transmission-frequency characteristics are therefore measured with this accuracy.

Single-control setting of the center frequency is provided by mechanical coupling of the klystron cavity tuning and repeller-voltage controls. Frequency is identified by the resonant-cavity frequency meter, a diode rectifier within the meter producing a rectified pulse as the output frequency of the klystron traverses the resonant frequency. This pulse, after amplification and clipping, is added directly to the oscilloscope signal trace. The frequency-identifying pulse is not transmitted through the circuit under test. Therefore, "stripping" of the pulse when measuring through limiters is avoided. Also, instant assurance that the RF generator is operating properly is provided when, due to a trouble condition in a circuit under test, no test signal is received from it. The frequency meter, a TE_{011} mode resonator having a loaded Q in excess of 8000, is directly calibrated in 1-mc increments. Without correction (for initial calibration error, scale error, temperature and humidity) any frequency from 5850 mc to 6500 mc can be measured with an accuracy of ± 2 mc. Of more significance in determining band-edge responses of RF-RF and RF-IF circuits in the system, frequency differences up to 125 mc are measured with an accuracy of ± 0.5 mc.

3.6 *IF Generator*

The IF generator is functionally analogous to the RF generator, but, due to the lower operating frequency, employs lumped-element and coaxial-line circuitry. The IF test signal is generated by a balanced inductive-feedback oscillator. The frequency is swept over the IF range by varying the current through a saturable inductor in the "tank" circuit. A high-gain, fast-acting automatic level control, which acts on the anode supply voltage of the oscillator tubes, maintains the output at $+20 \pm 0.05$ dbm as the oscillator sweeps over its frequency range. A well-matched source impedance (a Thevenin generator) results from a series resistance of 75 ohms which is in the center conductor of the coaxial output beyond the sampling point for automatic level control. Filtering and balance maintain the harmonic output 40 db or more below the fundamental. The four-dial decade attenuator shown in Fig. 3 adjusts the output power in 0.1-db steps between $+20$ dbm and -71

dbm. The residual frequency characteristic, with the IF detector directly connected to the attenuator output, is less than ±0.05 db over the IF band.

The center frequency can be set at any point in the IF band. Frequency is marked in the same way as in the RF generator, but a lumped-element LC resonator is used. The resonator has a Q of about 200, is directly calibrated at 1-mc intervals, and permits identification of frequency with an accuracy of ±0.2 mc.

3.7 *RF Detector*

RF detectors, having the internal configuration shown in Fig. 5, are used as the receiving elements in RF-RF and IF-RF transmission measurements, and as the sensing element in the automatic level control of the RF generator. Since flat response for every silicon diode is essential and high sensitivity is not a requirement, variations in diode

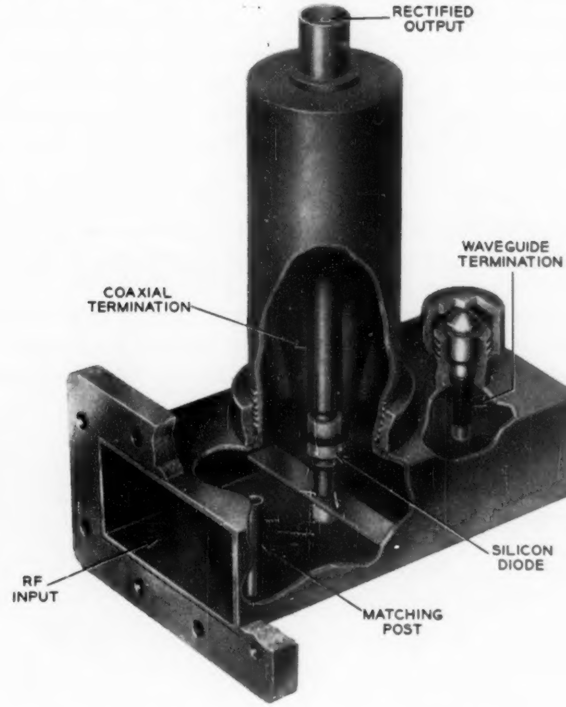

Fig. 5 — Internal configuration of the RF detector.

parameters are masked by a coaxial resistor connected effectively in series with the diode. Further stabilization of the input impedance is obtained by ending the waveguide in a well-matched resistive termination instead of the usual short circuit. A metal post (inductive reactance) completes the impedance-matching structure. Experimental determination of the optimum mechanical dimensions for these elements resulted in an input reflection coefficient of less than 0.14 (>17-db return loss). The response is constant to ±0.05 db over any 32-mc segment of the band.

3.8 *IF Detector*

The receiving element for IF-IF and RF-IF transmission measurements is the IF detector pictured in Fig. 6. The same circuit is used as the sensing detector in the automatic level control of the IF sweep generator. As in the RF detector, rectification is effected by a silicon diode. A rectifier load of low, predominantly resistive impedance causes the detector to respond to the half-wave average, rather than the peak value, of the input signal. This reduces errors due to noise and to even-order harmonics of the test signal. The rectification efficiency of the IF detector is constant within ±0.05 db, and the reflection coefficient is less than 0.032 (>30 db return loss) over the IF band.

3.9 *RF Reflectometer*

The waveguide components of the RF reflectometer are identified on the photograph of Fig. 7. The hybrid junction is a conventional

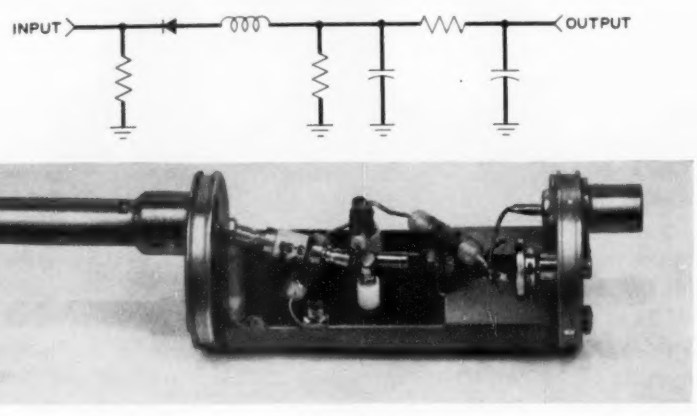

Fig. 6 — The IF detector.

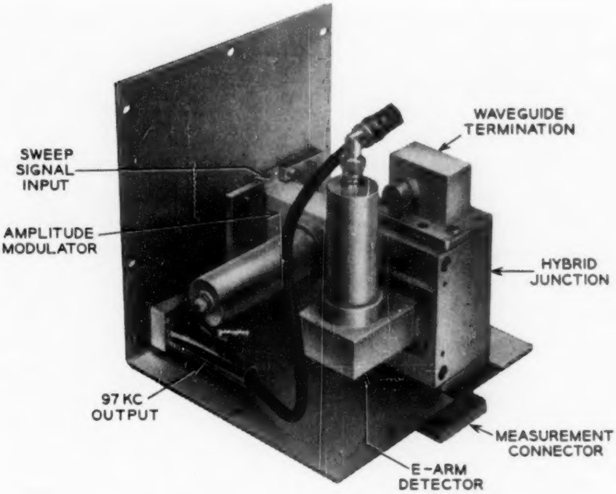

Fig. 7 — Waveguide components of the RF reflectometer.

E-H Tee waveguide junction, containing matching structures to give a well-matched impedance at all of its four waveguide inputs. Mechanical symmetry can be counted on in hybrid junctions of this type for achieving a very high (45-db) balance between the E and H arms. The detector connected to the E-arm of the hybrid has an internal configuration very similar to the one shown in Fig. 5 except that, to increase the sensitivity and hence signal-to-noise ratio, the termination has been replaced by a short circuit located approximately one-quarter wavelength from the silicon varistor. Essentially the same structure is also used in the amplitude modulator, which is connected between the H-arm of the hybrid and the sweep signal input; in this case the waveguide, instead of being terminated, ends in a waveguide flange which attaches to the H-arm.

The hybrid junction compares impedances connected to the measurement connector with the waveguide termination connected on the opposite arm. This termination, a carbon deposited resistor matched to the waveguide in a structure identical to that shown in Fig. 5, has a reflection coefficient below 0.01 over the RF band. This characteristic essentially sets the residual error in reflection coefficient measurements at ± 0.01; residual unbalance in the hybrid junction and inaccuracies in the attenuator, the (square) law of the silicon diode, and the standard waveguide load introduce errors that are small in comparison with

0.01.* The standard waveguide load, which is connected to the measurement connector for the initial setting of sensitivity, is just like the waveguide termination; however, the resistor has a resistance approximately 14 per cent higher to give a calibrating reflection coefficient of 0.178 (15-db return loss).

Use of 97-kc amplitude modulation on the swept signal input and subsequent demodulation, amplification and rectification of a 97-kc signal avoids the need for a high-gain dc amplifier. Internally generated noise in the E-arm silicon diode places a limitation on the lowest reflected signal levels that can be accurately measured. A bandwidth of 3 kc in the 97-kc amplifier is sufficient to display, with little error, the most rapid variations in reflection coefficient that are measured in the system. This selectivity, a low-noise 97-kc amplifier design, and maximum sensitivity in the E-arm detector (consistent with an adequate frequency response) results in a signal-plus-noise display which differs from that due to the signal alone by only 0.3 db. This performance applies when the impedance under test is excited with an incident wave of −1 dbm and has a reflection coefficient of 0.0178 (35-db return loss). The same noise performance would apply at −11 dbm and 25-db return loss, etc. Under these conditions (and at higher levels or higher reflection coefficients) uncertainty in the measured result due to noise is small.

3.10 *IF Reflectometer*

The IF reflectometer employs the same basic principles as its RF counterpart, but differs in instrumentation techniques because of the lower operating frequency. Fig. 8 shows the IF reflectometer, which is a simple equal-arm resistance bridge with a self-contained silicon diode detector. The IF generator, amplitude modulated at a 97-kc rate and simultaneously frequency-swept, supplies the test signal. A 97-kc regulating loop, acting on the oscillator of the IF generator, maintains the modulation constant as the carrier frequency is swept. The 97-kc amplifier-rectifier and indicating circuits are those used for the corresponding RF measurements; the calibrating and measuring procedures are exactly the same.

Noise generated in the diode detector places a limitation on the lowest reflected signal levels that can be accurately measured; for

* Field experience, in connection with measuring the return loss of long waveguide runs, shows that even better performance can be obtained by using a more precise reference termination, of 60-db or better return loss, and by using isolators to suppress secondary reflections, e.g., between the hybrid junction and the amplitude modulator.

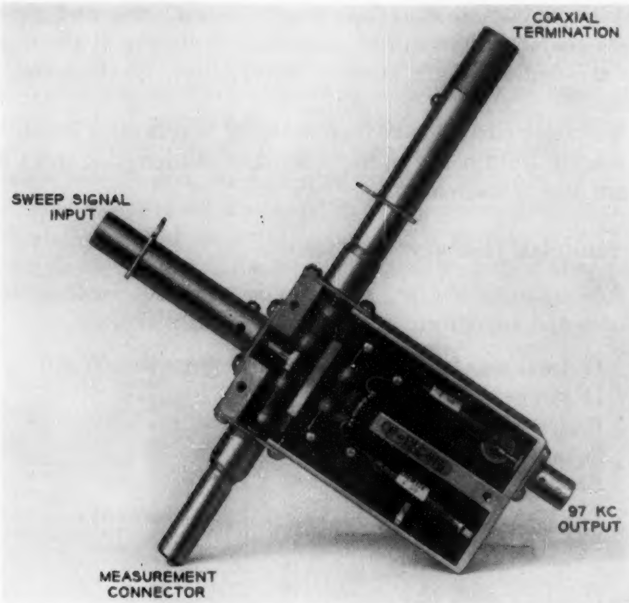

Fig. 8 — The IF reflectometer.

equivalent performance, the power in the impedance under test must be kept 3 db higher than in the RF case because of the added loss in the resistance bridge. The residual error in reflection coefficient is essentially set at ± 0.01 by the return loss characteristic of the coaxial termination, as the resistance bridge contains precision resistors and capacitance trimmers that permit realization of a 50–60-db balance over the IF band.

3.11 *Indicating Circuits*

Transmission and reflection coefficient measurements are presented as traces on the screen of a five-inch oscilloscope especially designed for this application. In this instrument conventional direct coupled amplifier circuitry is employed, but internal sweep generating circuits are omitted, the sweep voltages being supplied by the sweep generator of the test set.

As an operating convenience, a dc potential of adjustable magnitude is substituted for the signal during the return trace. This appears as a

horizontal reference line against which transmission and reflection coefficient characteristics can be compared. Switching of the reference trace is accomplished by a mercury relay, driven synchronously with the sweep.

The indicating circuits have been designed to provide a sensitivity of approximately 1 db/in. for all transmission measurements, and ½ db/in. for return loss measurements.

IV. FM TERMINAL TEST SET

The FM terminal test set provides measurement facilities for the surveillance and adjustment of the following quantities:

IF transmission and reflection coefficient (50–100 mc)
IF power and frequency (74.13 mc)
Baseband transmission (0.1–10 mc)
Square wave response (62 cycles)
FM receiver deviation sensitivity
FM transmitter deviation and optimum linearity
FM receiver linearity

4.1 *Some General Design Features*

To make all the required measurements with a single mobile console, the measurement methods make multiple use of as few basic circuit units as possible. In this set, shown in Fig. 9, all of the measurements are accomplished by functional rearrangement of 17 basic circuit units. The functional changes are made with switch controls and patching plugs in the jack field. All connections to the circuit under test are made with flexible cables from jacks on the front panel, using 75-ohm coaxial for IF and unbalanced video signals and 124-ohm shielded pair for balanced video signals.

Wherever feasible, the same features which contribute to accuracy, convenience, and speed of measurement and to internal accessibility for repair and adjustment in the transmitter-receiver test set have been incorporated in the FM terminal test set.

4.2 *IF Transmission and Reflection Coefficient*

The measuring methods employed for sweep frequency IF transmission and reflection coefficient measurements are identical to those described for the transmitter-receiver test set; circuits and instrumentation techniques are also substantially identical.

4.3 *IF Power and Frequency*

A silicon diode rectifier in a circuit like that of the IF detector (Fig. 6), but with an input impedance-matching network and an input attenuator, forms the IF power meter. The matching network increases the sensitivity but restricts the bandwidth to ± 0.5 mc around 74.13 mc. The range is -15 dbm to $+15$ dbm with an accuracy of ± 0.5 db; indication is on one of the scales on the dc meter shown in Fig. 9.

A crystal-controlled oscillator, generating a frequency of 74.130 ± 0.003 mc, is used to check the carrier frequency of the FM transmitter by a comparison method. This oscillator is also used as a signal source

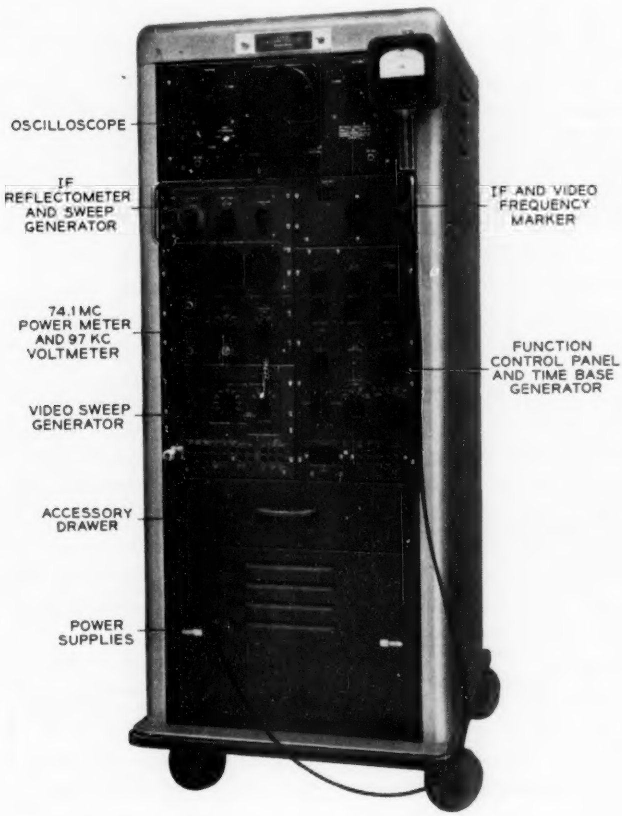

Fig. 9 – Front view of the FM terminal test set.

for the adjustment of FM receiver carrier balance and in the video sweep generator described below.

4.4 *Video Transmission*

For 0.1–10 mc video transmission measurements a sweep-frequency technique is employed. The functional diagram of Fig. 10 shows the method of measurement to be similar to that described for RF and IF transmission. The video test signal is generated as the difference frequency obtained by applying to a modulator the swept frequency from the IF sweep generator, together with the 74.13-mc signal generated by the crystal-controlled oscillator. Sweep range and centering controls permit adjustment of the IF generator to sweep over the range 74.3 mc to 84.3 mc; the resultant baseband sweep covers 100 kc to 10 mc. Fast-acting automatic level control, together with appropriate equalization, maintains the baseband test signal constant at +10 dbm within ±0.01 db as measured through twelve feet of flexible balanced cable.

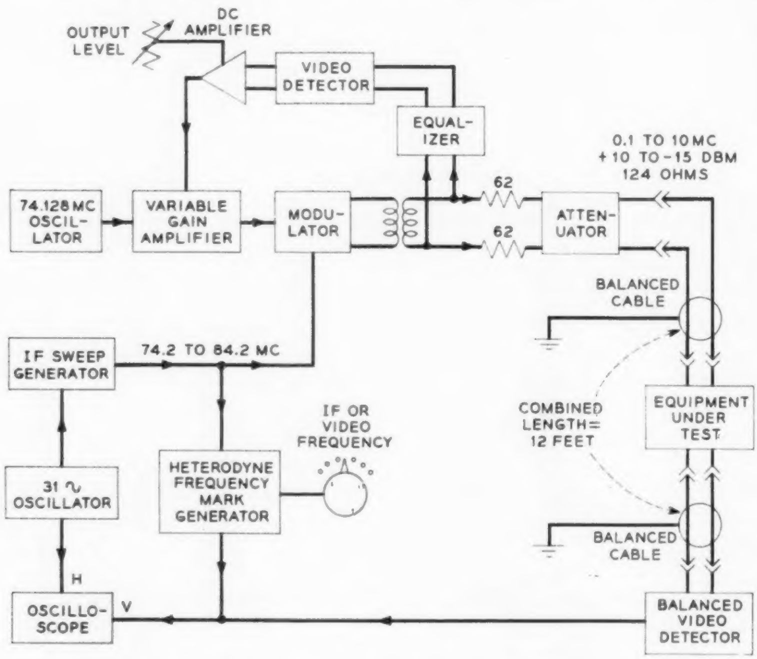

Fig. 10 — Block diagram for video transmission measurements.

Thus no error is incurred so long as two cables of the same total length are used to interconnect between the test set and circuit under test, as shown in Fig. 10. The output impedance is 124 ohms balanced. The balanced attenuator, 0–25 db in 0.5-db steps, permits selection of output levels from +10 to −15 dbm. Balanced and unbalanced video detectors are provided; the former, schematically shown in Fig. 11, employs a full-wave bridge rectifier with low impedance load to ensure immunity to noise, harmonics, and longitudinals. The unbalanced baseband detector circuit is similar to that of the IF detector, but with element values appropriate to the lower frequency range.

The residual frequency characteristics of less than ±0.05 db for the combination of generator, output attenuator, and video detector allow video transmission to be measured with this accuracy over the 10-mc

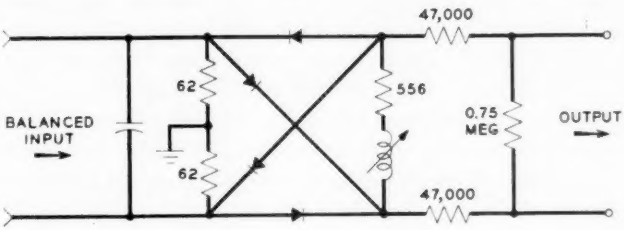

Fig. 11 — Circuit of the balanced video detector.

band. The displayed characteristic on the oscilloscope can be resolved to better than ±0.01 db.

A heterodyne receiver, having a 10-mc intermediate frequency with a 100-kc passband, is used to generate frequency identification marks on the transmission display. A resolution of ±0.1 mc and an accuracy of ±0.25 mc is obtained. This unit, permanently connected in the IF sweep generator circuit, is used for both video and IF identification, the dial having separate scales for the two applications.

4.5 *62-cps Square-Wave Test*

A simple square-wave test, at a frequency of 62 cps, serves to expose irregularities in low-frequency transmission (amplitude and phase) of the FM terminals which might impair television transmission. The test signal is produced by a relay, driven at 62 cps, which alternately makes and breaks a connection to a dc source. By comparing the response with illuminated horizontal scribe lines displayed on the oscilloscope screen,

deviations from an ideal flat-top, square-wave response are measured as a percentage of the peak-to-peak amplitude. Measurement accuracy, limited mainly by resolution of the oscilloscope scale, is better than $\frac{1}{2}$ per cent.

4.6 *FM Receiver Deviation Sensitivity*

Fig. 12 shows the method for setting, to a specified value, the sensitivity of the FM terminal receiver to a reference frequency deviation. A synthetic square-wave frequency modulated test signal is created by alternately keying on and off two crystal-controlled oscillators, one having a frequency 1.98 mc higher, the other 1.98 mc lower than 74.13 mc. The keying rate is 97 kc. The receiver output, under these conditions, is a 97-kc square wave. An extremely stable, selective, back-biased, electron tube voltmeter is provided to measure the fundamental component of this square wave. Since the relationship between the amplitude of a square wave and its fundamental is known, the sensitivity of the receiver is thus established. By appropriate selection of the keyed oscillator frequencies (76.112 and 72.147 mc), and of the values of attenuators and resistive coupling networks, need for computation on the part of the operator is eliminated; a "zero" indication of the voltmeter establishes that the receiver sensitivity has been properly adjusted.

Zero indication occurs when the dc output of the electronic voltmeter equals a potential determined by a voltage-reference gas tube. Whenever the gas tube is replaced, the over-all sensitivity of the voltmeter for this zero indication must be readjusted, by comparison with a known 97-kc voltage. Subsequent variations in voltmeter sensitivity are corrected by a self-calibrating feature illustrated in Fig. 12. Prior to each measurement a regenerative loop through the voltmeter and back through stable, fixed attenuators is established by closing the calibrating switch. The gain calibration control is then adjusted so that oscillation at 97 kc is barely sustained. By this method the gain through the active circuit is made equal to the loss through the feedback path attenuation, which is accurate to ± 0.01 db.

As an end result, the long-term accuracy of the voltmeter, mainly dependent upon its initial calibration and the stability of the gas tube reference voltage, is ± 0.05 db. Since the ± 1.98-mc frequency deviation, generated by the keyed crystal oscillators, is accurate to one part in 10^3 (0.01 db), it is expected that receiver sensitivity adjustments can be gauged with an accuracy of better than ± 0.06 db.

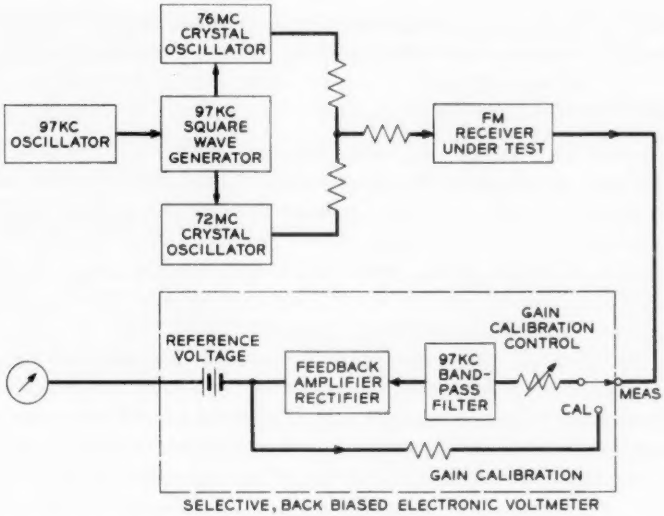

Fig. 12 — Block diagram for FM deviation sensitivity measurement.

4.7 *FM Transmitter Deviation and Optimum Linearity*

Frequency deviation of the FM terminal transmitter is adjusted by applying a 97-kc sinusoidal input signal to the transmitter and setting to a specified value the resulting frequency deviation as observed with an FM receiver. This FM receiver is a specialized one, is an integral part of the test set, and is calibrated just prior to use by the technique and instrumentation of the previous section. The amplitudes of the sinusoidal signals at the transmitter input and receiver output are measured sequentially with the selective voltmeter. By the use of suitable values of fixed attenuators and resistive coupling networks, properly adjusted deviation is indicated as a "zero" (actually, mid-scale) reading on the dc meter of the voltmeter. Accuracy of the deviation adjustment will be very nearly the same as for FM receiver sensitivity — better than ± 0.06 db.

Optimum linearity of frequency modulation occurs at the repeller voltage for the DO klystron in the FM transmitter, for which small signal deviation is a minimum.[2] Using this principle, optimum linearity is obtained during the adjustment of transmitter deviation, by setting the repeller voltage for a minimum reading on the dc test meter. Differential sensitivity of the electron tube voltmeter ($\pm\frac{1}{16}$ inch deflection

for ±0.01 db of level change) is adequate to assure an optimum setting well below linearity requirements set by system considerations.[2]

4.8 *FM Receiver Linearity*

FM receiver linearity (or, more properly, nonlinearity) is defined as the percentage change in the small-signal FM sensitivity (volts/mc) of the receiver over the IF band. Linearity is measured by a method that was first described in connection with TD-2 radio relay system.[4]

A low level, 97-kc signal, producing a deviation of about ±200 kc, is applied to the repeller of the BO klystron of the FM transmitter. Simultaneously, the transmitter's output frequency is deviated over the range 58 mc to 90 mc by application of a 31-cps high-level signal to the DO klystron. The resultant composite FM signal is applied to the input of the FM receiver under test. The same amplifier-rectifier that is used for return loss measurements is now used for selection, amplification, and envelope rectification of the 97-kc component of the receiver output. Variations in this rectified output, displayed on the oscilloscope screen, are measured by comparison with an illuminated horizontal scale having lines which are spaced 0.2 inch apart. Vertical sensitivity, calibrated with the return loss measuring attenuator contained in the amplifier-rectifier unit, is normally set for 0.2 db (approximately 2 per cent) per scale division. Instantaneous frequency is identified by placing, on the display, marks generated by the heterodyne frequency mark generator described above in Section 4.4.

By separately modulating the two klystron oscillators of the transmitter, effects of nonlinear modulation at 97 kc are avoided. The accuracy of measurement is therefore primarily limited by scale resolution and accuracy of the calibrating attenuator; major importance is attached to linearity in the 64–84-mc range, for which nonlinearity is ordinarily less than ±5 per cent. An accuracy of $\pm\frac{1}{4}$ per cent is obtained for such measurements.

V. PROTECTION SWITCHING TEST SET*

This test set is designed for routine tests on the operation of the protection switching system,[3] and for confirmation of adequate performance after repairs have been made. It also provides mounting and testing facilities for individual circuit units that have been removed from the system for repair or adjustment.

In contrast to the analog signal testing techniques employed in the

* H. I. Maunsell made important contributions to this section.

previous two instruments, the protection switching test set features mostly digital testing techniques. While high accuracies of measurement are not required, ability to test the system operation and to make a large number of test observations on as many as 50 different kinds of circuit units is provided. A photograph of the protection switching test set is shown in Fig. 13.

5.1 *System Operation Tests*

The common control panel, identified on the photograph, is connected to the transmitting logic of the protection switching system by connectors on a multi-conductor cable. Digital signals are applied to the logic by manual operation of keys on the control panel. Operation of these "manual exerciser" keys, in a systematic sequence, simulates to the system's logic a failure in the regular broadband radio channels, one at a time. Protection switching action is thereby initiated, and lamps in

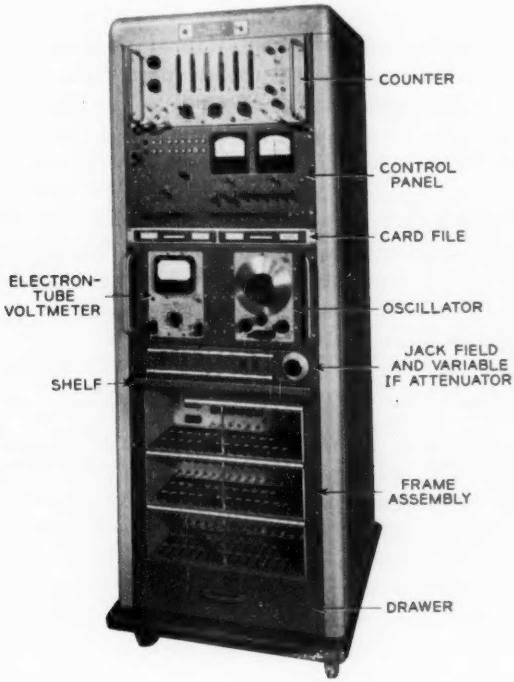

Fig. 13 — Front view of the protection switching test set.

the test set (and in the TH system) indicate the system's response to these simulated failures.* When the response is abnormal, the lamp indications serve to localize where the trouble exists to within three or four circuit units in the system. These units, all or one at a time, are replaced with spares, the logic exerciser being used to confirm that normal operation has been restored.

The time interval between initiation of a simulated failure and completion of protection switching action can be measured with the electronic counter identified on the photograph. Similarly, time intervals between input and output events, important in some of the individual circuit units, can be checked with the counter which is a standard commercial design.

5.2 *Circuit Unit Tests*

Any individual circuit unit can be tested by inserting it in the proper one of the 48 slots in the frame assembly. Insertion of the circuit unit applies power to the unit and connects it to the appropriate testing circuit and resistance termination. Manual operation of the "unit testing" keys on the control panel applies appropriate dc signals (codes) to the unit's input terminals. The response to these signals is indicated on the dc meters. Correlation of the sequential responses provides an aid to localization of trouble in the circuit when these responses are abnormal.

This multi-receptacle method for testing circuit units materially reduces setup time. Testing speed and convenience are also increased by including on the test set a card file giving information (in the form of tables for each type of circuit unit) on key operating sequences and the normal meter responses.

5.3 *Analog Signal Tests*

Included in the test set are a low-frequency electron tube voltmeter and oscillator of standard commercial design. These instruments and the electronic counter are used to check and adjust frequency, voltage output and frequency response of tone transmitter and tone detector circuit units of the system.

To measure powers of around −21 dbm at the output of 74.1-mc IF carrier supply units, a silicon-crystal rectifier is included. The rectifier input appears on the jack field, and its output is applied to one of

* Currently under development is an "automatic exerciser" which will do this regularly once a day and initiate an alarm in case of abnormal response.

the dc meters. A continuously variable (stepless) 5–15-db attenuator is also supplied, to permit transient-free variation of test IF signals at the input of IF level detector circuit units, when checking and adjusting the IF carrier level changes that will initiate protection switching action.

Video transmission, IF transmission and IF reflection coefficient are measured, when required, on certain of the circuit units while inserted in the frame assembly. These analog tests are made with an FM terminal test set; the IF tests can be also made with a transmitter-receiver test set.

VI. MAINTENANCE CENTER TESTING FACILITIES

Illustrated by the line drawing in Fig. 14 are the equipment mounting racks, power supplies and other facilities that are permanently installed in a maintenance center for repair and testing of units that have been removed from the system for repair or adjustment. Not shown, but normally available for use at this center, are the three mobile test

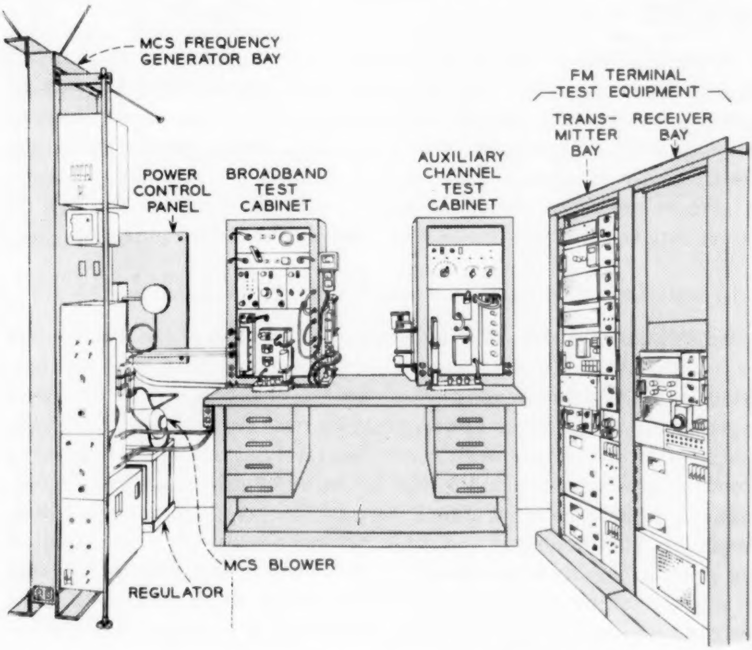

Fig. 14 — Line drawing showing test facilities at a TH maintenance center.

sets described above as well as the regular and special tools and test instruments that are essential at any electronic repair center for maintenance of radio relay equipment. For a 4000-mile system there may be only ten of these repair centers, since each may service system equipment in one or two main switching stations and up to twelve or more radio repeater stations.

As may be inferred from the illustration, the maintenance center is essentially a skeleton radio repeater, including a modified carrier supply and auxiliary channel radio equipment. Units of this maintenance center equipment may be replaced by units removed from the working system to adjust, test and confirm adequacy of repair operations on them. Extender cables permit access to normally covered areas of the unit under test.

Also installed in this center is a repackaged FM terminal transmitter and receiver into which plug-in units from a working system terminal may be connected, either directly or by extender cables.

VII. POWER SUPPLIES

Power supplies for the test equipment are described in some detail in a companion paper.[7] The mobile consoles operate from 117-volt ac convenience outlets, and the maintenance center from a permanently connected 208–230-volt line. On the consoles, the dc power supplies are electronically regulated; some critical heaters are supplied with dc. At the maintenance center the main ac line is regulated and the dc power supplies are the same as those used in the radio equipment.

VIII. STRAIGHTAWAY GAIN AND DELAY DISTORTION MEASUREMENTS

Adjustable gain and delay equalizers will be used in the TH System to compensate for (to mop up) residual distortions that have accumulated over an equalization section that may include up to ten radio repeaters.[8] Test facilities are required to measure and administer adjustment of these equalizers. Since the receiving station is separated from the test-signal sending station by many miles, straightaway measurement techniques are required. As a further requirement, continuous displays of gain-frequency and delay distortion-frequency characteristics are essential for any comprehensive adjustment of equalization intended to produce optimum performance of the system. With these continuous displays, an operator can adjust the equalization to minimize distortion in the end-to-end characteristic.

Included in each FM terminal test set are circuits which permit its

use at the sending and receiving stations for sweep-frequency gain measurements. At the sending station, the sweep-frequency video output is combined with a low-level sample of the 31-cps voltage from the time base oscillator (which is sweeping the video frequency). This combined signal is applied to an FM terminal transmitter and sent over the radio link.

At the receiving station, an FM terminal receiver returns the test signal to baseband frequencies. The 31-cps signal is recovered in a separation network, amplified and applied to the time base oscillator circuit in the FM terminal test set at the receiving station. This oscillator's frequency is thereby forced into synchronization with the distant sending time base oscillator. Consequently, the displayed characteristic will remain stationary, since the horizontal deflection for the receiving oscilloscope will be synchronized with the sending video signal sweep.

An adjustable 31-cps phase shifter at the receiving station permits centering the display, and resonance transmission markers at 5 mc and 10 mc provide instantaneous received frequency identification.

There is already available for the TD-2 radio system a straightaway delay distortion test set in which the small-signal phase shift at 278 kc is measured while the center frequency is swept slowly, at 100 cps, over the IF band of interest.[9] The delay distortion-frequency characteristic is displayed on the screen of an oscilloscope. This same equipment can be used in conjunction with TH FM terminals for delay distortion measurements of the TH radio system.

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